

For the Above and Beyond Crowd

Here's a neat RF modulator/detector for use with surplus microwave SWR meters.

Microwave test equipment has never been plentiful or inexpensive. This fact becomes readily apparent as soon as you want to make a measurement of power, gain, frequency, and so forth. Here's something you can build for yourself.

The earliest microwave experimenters found value in modulating their sources with a 1 kHz tone and using a simple crystal detector followed by a tuned audio amplifier and voltmeter as a measurement tool. The technique was quickly adapted for use with slotted lines for making impedance and SWR measurements.

Even today, the technique is used for amateur radio VHF/UHF/Microwave antenna gain measurement contests. A tone-modulated oscillator or signal generator with a broad beamwidth antenna

is used as a source, and a crystal or video detector followed by an HP 415 or similar SWR meter is used to make power measurements relative to an antenna with known gain.

This idea works well if your source can be easily amplitude-modulated (AM). However, most microwave brick oscillators, klystrons, and Gunn diodes do not lend themselves to direct AM modulation. Most can easily be frequency modulated (FM), but AM must be done with an external modulator such as a PIN-diode. The only PIN-

diode switches I own exhibit a large reflected impedance difference between their on and off states which can upset the stability of some sources.

I wanted to be able to test with a variety of microwave sources without having to build a unique dedicated modulator for each source, so I decided to modulate the receiver instead of the transmitter. This idea is not that much different than what is done with a Dicke radiometer or even the earliest automotive radar detectors. The idea is to chop the input line to the detector. With no RF present, there is no output. When RF is present, the output is an AC signal at the chopping frequency, proportional to the amplitude of the incoming RF. The only difference between the two approaches is that with this system, the meter will respond to all RF present, whereas with the modulated source, the meter only responds to the modulated RF. While a simple RF bandpass filter should cure any interference problems if they occur, I have yet to need one.

Years past, at a hamfest, I had purchased several Hewlett-Packard (HP) PIN-diode SPDT switch modules as an investment that now seemed ideal for the task. All I would have to do would



Photo A. Front view of modulator/detector.

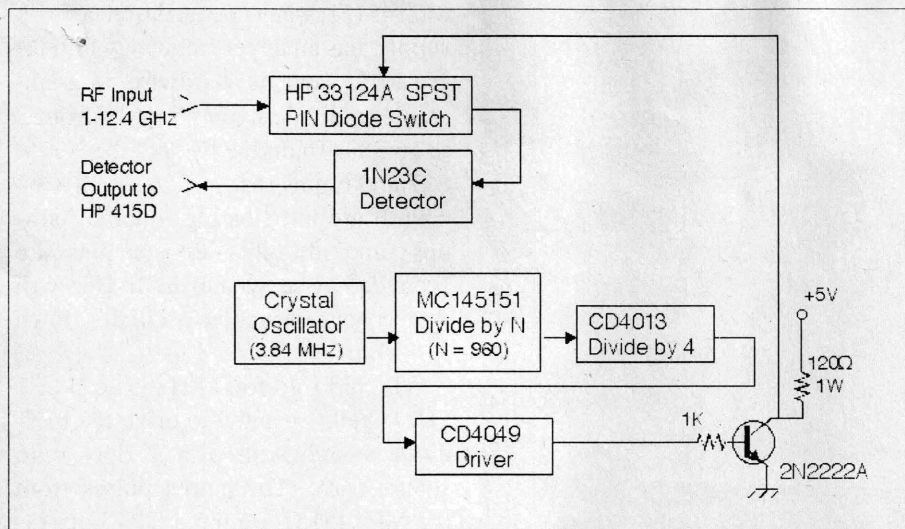


Fig. 1. Block diagram of HP 415D modulator/detector.

be to feed the received RF to one of the switches, turn the switch on and off at a 1 kHz rate, and follow the switch with a broadband RF detector. The result would be a universal adapter to allow me to use my HP 415D SWR meter with any source within the frequency range of the PIN-diode switch and detector. The PIN-diode switch I used, a Hewlett-Packard model 33124A, has a bandwidth of 0.1 to 12.4 GHz, and the detector I used, a Sage 1021H, has a bandwidth of 1 to 12.4 GHz. The net result is a system bandwidth of 1 to 12.4 GHz.

A block diagram of the system is shown in Fig. 1. A complete schematic diagram of the PIN-diode switch

driver is presented as Fig. 2. I have several HP PIN-diode switches, but the 33124A was the only one I could use in a negative ground system.

I could have used a simple, free-running, NE555 1 kHz oscillator to drive the PIN-diode switch, but decided instead to use a crystal-controlled driver for several reasons. First, the cost of a crystal oscillator and divider chain is not that much more than a free-running oscillator. Second, a very stable source would be needed to take full advantage of the fact that the HP 415 bandwidth can be narrowed down from 100 to 15 Hertz. And, third, I like the idea of no

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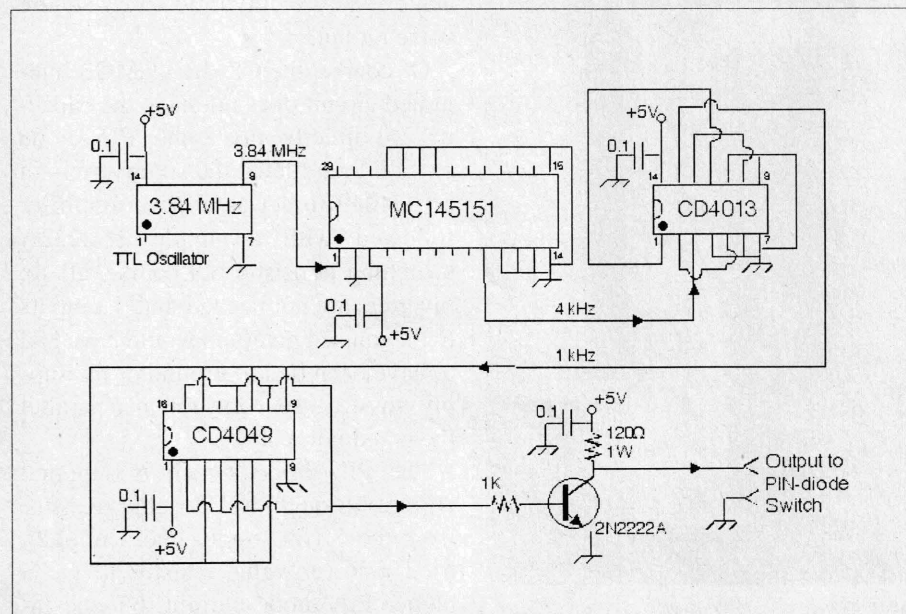
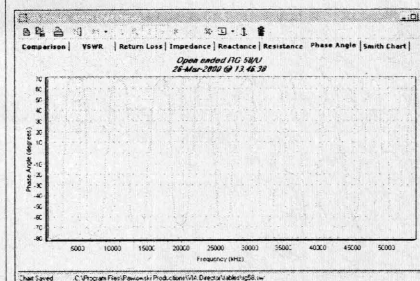


Fig. 2. Schematic of 1 kHz generator and driver circuit.

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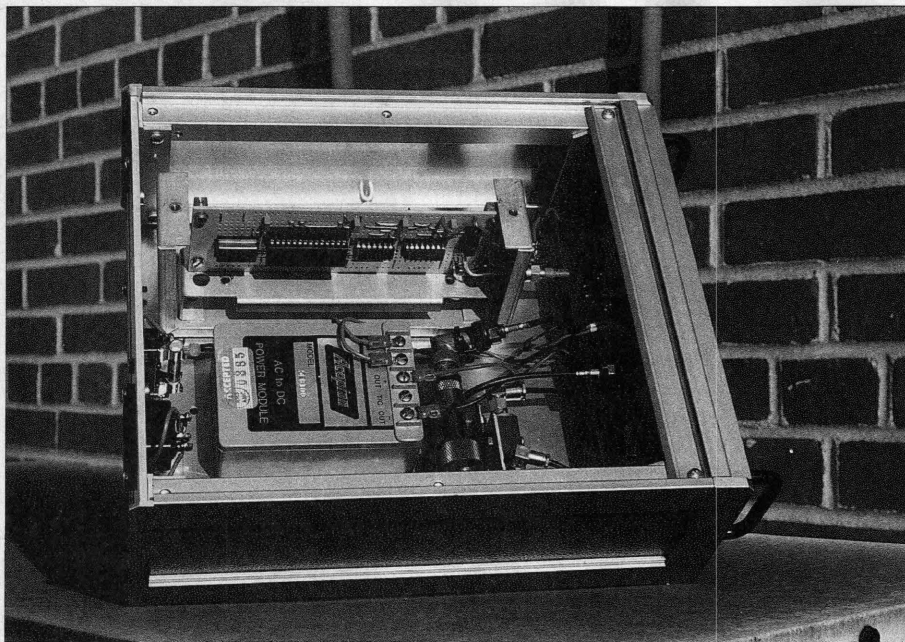


Photo B. Modulator/detector with cabinet cover and driver shield removed.

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adjustments or calibration required — too many knobs spoil the measurement.

Having decided to start with a crystal oscillator, why didn't I use a common 1 MHz TTL oscillator? Well, a quick check of my junk box revealed several TTL unit oscillators. I hated to use one of my 1.000 MHz units when I had some other units with seemingly odd-ball frequencies. I selected a 3.84 MHz unit

because it will probably otherwise never be used. Use whatever oscillator you have, as long as the output frequency is an even integer multiple of 1 kHz.

Following the oscillator, I used an MC145151 integrated circuit to divide the 3.84 MHz by 960 to get 4 kHz. The MC145151 was designed to be part of a phase locked loop and has many internal functions which are not needed for this application. However, the divide-by-N portion can be used independently of the other functions. The

MC145151, when operated from a 5 volt supply, has an upper frequency limit of 15 MHz and is hardwire-programmable to divide by any integer from 3 to 16,383. To divide by 960, the N = 9, 8, 7 and 6 pins ($512 + 256 + 128 + 64 = 960$) are left floating (internal pull-ups) and all other divisor pins are grounded. Use whatever divisor will divide your particular oscillator down to 4 kHz.

Why did I go for 4 kHz instead of 1 kHz? Well, I wanted to drive the PIN-diode switch with a 50% duty ratio square wave. The output pulses from the MC145151 are not a 50% duty ratio square wave, but have a pulse duration equal to the period of one cycle of the input signal, 0.26 microseconds in this case. Therefore, I knew I would need to add at least one binary flip-flop following the MC145151 to achieve the 50% duty ratio. Since most integrated circuit flip-flops have at least two stages per package, the output of the MC145151 could be either 2 or 4 kHz. I chose 4 kHz. The divider I selected to follow the MC145151 is a CD4013, a common two-stage type D flip-flop. A type D flip-flop will toggle or divide by two if the set and reset pins are grounded and the not-Q (complementary) output is connected to the D input. The flip-flop input is the C or clock pin, and output is taken from the Q pin. As I mentioned, I used both stages to divide by four and achieve a 50% duty ratio 1 kHz square wave output.

Of course, the CD4013 CMOS integrated circuit does not have the capacity to directly drive the PIN-diode switch. I paralleled the six inverters in a CD4049 to act as a drive amplifier followed with a simple 2N2222A switching transistor. Of course, all six inverters are not needed, but the inputs to the unused inverters would have had to have been tied to ground or the supply anyway, so why not just parallel them and not worry.

The PIN-diode switch is supplied current through a 120 ohm resistor from the 5 volt supply line. I initially tried a lower value resistor to get a higher PIN-diode current, but the increase in signal output was marginal,

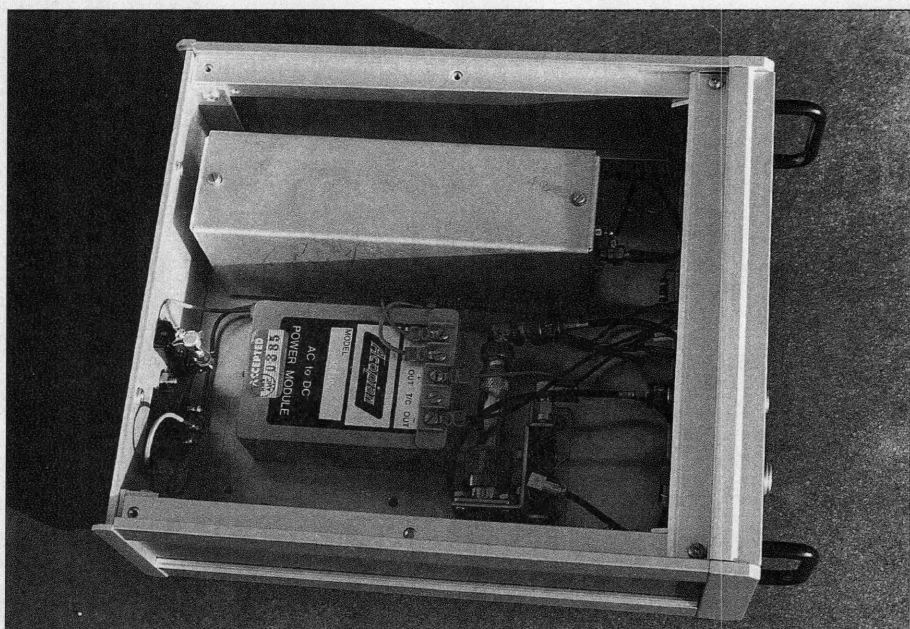


Photo C. Modulator/detector with cabinet cover removed. PIN-diode driver at top, power supply in the middle, and PIN-diode switch with detector in lower right.

and the power consumption and radiated noise level were much higher. The difference in detected output between using a switch that switches from 0 to 10 dB versus one that switches from 0 to 100 dB is less than 10%.

The 2N2222A transistor, when switched off, has no influence on the PIN-diode current. When switched on, the 2N2222A shunts all the current through the 120 ohm resistor to ground, thereby reducing the PIN-diode current to zero. The PIN-diode switch is on when the transistor is on and vice versa. By using a shunt switching configuration, the current through the 120 ohm resistor is relatively constant. This way, no large switching currents are generated that can radiate and be picked up by the HP 415D SWR meter. After all, the HP 415D has a useful sensitivity approaching 0.1 microvolt, and the final system noise floor will determine the ultimate sensitivity of the complete instrument.

The voltage across the PIN-diode switch is 0.1 V when the transistor is on and 0.9 V when the transistor is off and all of the current is passing through the PIN-diode. The 120 ohm current limiting resistor is dissipating 0.2 W when the transistor is on and 0.14 W when the transistor is off. The average power dissipated by the resistor is 0.17 W. For reliability, I used a resistor with a 1 W rating. The current drain on the 5 V regulated power supply, not including the LED power indicator, is approximately 60 mA. The LED power indicator draws an additional 12 mA.

To further reduce radiated noise, the driver circuit board is mounted inside a shielded enclosure and the TTL oscillator module and each integrated circuit is bypassed with a 0.1 μ F capacitor. The circuit board, mounted in its shielded enclosure, is shown in the photographs both with the shield removed and in place.

The shielded driver is mounted in a larger aluminum cabinet along with the regulated 5 volt power supply, PIN-diode switch, and crystal detector. The cabinet volume is less than the HP 415D and is easily stacked under or

over the HP 415D as seen in the photographs. Coax cables are used internal to the cabinet between the front panel feedthrough connectors and the PIN-diode switch input and the detector output. A short SMA jumper cable connects the PIN-diode switch and the detector as seen in the photographs. An SMA to SMC cable is used to route the 1 kHz drive to the PIN-diode switch. I mounted the PIN-diode switch and the detector on a common bracket for ease of assembly and to minimize the length of the interconnect cable. The usual paint trim, handles, and label decals were applied to the front panel.

In operation, the HP 415D "INPUT SELECTOR" switch is set to "XTAL," either 200 ohm or 200k depending on which position gives the strongest reading with your detector. I usually use the 200k position. Do not use the "BOLO" or bolometer position, as it is designed to supply a current bias to the detector and can destroy the crystal diode. If you use a Hewlett-Packard detector with its own, matched, square-law optimized load resistor, use the 200k setting.

Because a diode detector, when used with low level signals, is a square-law device, the dB scales are reasonably accurate relative power indicators. For example, a change in meter reading, up or down, of 3 dB, represents a factor of two power change. The unit can be used for antenna gain measurements, relative to a known antenna, attenuation measurements, and small signal amplifier gain measurements, as well as slotted line impedance measurements as explained in the HP 415D manual. While the unit was designed and built with the HP 415D in mind, it should work equally well with other 1 kHz tuned SWR meters having a crystal detector input.

How well does it work? I continue to be pleased with the performance and use it regularly. Using a mid-value gain setting on the HP 415D, the noise floor is mid-scale on the 50 switch position. At 1 GHz, a signal of -42 dBm is 3 dB (full scale deflection) above the noise floor. At 2 GHz, a signal of -47 dBm is also 3 dB above the noise floor. The only regret I have is that I did not build it sooner. 73

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Simple Band-Pass Filters for Receiver Projects

One approach to designing simple radio receivers is to use a broadband front end. Although at one time the front ends of radio receivers were tuned to a specific frequency, for the past two decades or so a broadband approach has been popular. The need for broadband tuning became apparent to me when I was working on some direct conversion receiver (DCR) designs. The DCR is similar to the superheterodyne (Figure 1), except that the local oscillator (LO) operates on or quite near the desired RF frequency.

For example, if you wanted to tune a DCR to receive a 7050 kHz CW signal, and wanted about a 1000 Hz beat note in your earphones, then the local oscillator would be set to either 7049 or 7051 kHz. When the local oscillator signal and the 7050 kHz RF signal is heterodyned together, the result is the

1000 Hz difference signal. An audio low-pass filter, or a transformer that acts inherently as a "low-pass filter" when the other signals in the pass-band are RF, at the output will select the difference signal over the LO and RF components that survive the heterodyning process.

Most DCRs use simple double-balanced mixers (DBM) of the diode variety, or Wilson Transconductance Cells (of the NE-602 IC variety) as the frequency converter. In both cases, strong signals appearing at the input can wipe out the effectiveness of the mixer. In addition, there are a number of other ways that even relatively weak out-of-band signals can appear in the output of the DCR. Although specific frequency tuning is used, it adds one tuning control. The minimalist design approach requires a bandpass filter at the front end that admits only those signals that are desired.

Figure 2 shows the circuit to a bandpass filter made by cascading low-pass and high-pass filter sections. In practice, some isolation may be required between the LPF and HPF sec-

tions, unless there is a relatively wide separation between the cut-off frequencies. Each section is a five-element design with 0.1 dB ripple (more or less), as defined in recent editions of *The ARRL Radio Amateur's Handbook*. The values for the components are found as follows:

$$\begin{aligned} L1 &= 9.126/F \\ L2 &= 15.72/F \\ L3 &= 9.126/F \\ L4 &= 5.803/F \\ L5 &= 5.803/F \\ C1 &= 4365/F \\ C2 &= 4365/F \\ C3 &= 2776/F \\ C4 &= 1662/F \\ C5 &= 2776/F \end{aligned}$$

where F is the frequency in megahertz (MHz), the inductances are in microhenrys (μ H) and the capacitances are in picofarads (pF).

The printed circuit board (PCB) pattern in Figure 3 is designed for miniature 7mm and 10mm coils of the sort made by Toko, and sold through Digi-Key (P.O. Box 677, Thief River Falls MN 56701-0677; 1-800-344-4539). However, if a toroidal core coil is used, which you can home-brew, then use the two outer pins on the three-pin side of the coils' foil pattern to make the connection (that corresponds to the internal wiring of the Toko coils). If you want a copy of the PCB, they are available for \$4.25 plus \$1.50 S & H

per order from FAR Circuits (18N640 Field Court Dundee IL 60118).

Figure 4 shows one version of the circuit intended for use at the front end of a high performance AM broadcast band receiver that I am working on. The filter has -3 dB cutoff points of 500 kHz and 2,000 kHz with the values shown. The completed filter, using the PCB of Figure 3 and the values shown in Figure 4, is shown in Photo A. The shielding is made from strips of 1" brass sheet metal strips. These strips are typically available in hobby shops that deal with model builders, as well as "lapidary" and amateur jewelry makers' shops. The sides of the box were bent square, then soldered. The corners were joined with soldered bits of brass right angle stock (available from the same sources).

By the way, if you wish to experiment with direct conversion receivers, you might want to look into the NE-602 IC and the Mini-Circuits passive double balanced mixers. Ocean State Electronics (P.O. Box 1458, Westerly RI 02891; 401-596-3080) lists one of the Mini-Circuits DBMs in their catalog. They also sell the MAR-x series of Mini-Circuits monolithic microwave integrated circuits (MMICs). Write for their 1994-95 catalog—it's interesting for the ham builder.

If there is enough interest, then I'll cover the basic direct conversion receiver designs in this column in the near future.

Book Note

Recently one of my publishers, Hightext Publications (P.O. Box 1489, Solana Beach CA 92075; 1-800-247-6553) sent me a newly-released book that I found thrilling. *Modeling Engineering Systems* by Jack W. Lewis is a real must for anyone interested in writing PC-based software models of electrical circuits and mechanical devices. This text grew out of a course the author took at MIT, as well as his own professional experience. Expecting to find this topic a real snooze, I was pleasantly surprised for Lewis' book makes the once-arcane art of modeling accessible to anyone with a basic math background.

Although at least introductory calculus is needed to get the most out of this book, calculus is not strictly necessary, as the clear, well-thought-out writing style makes complex concepts easily understood.

Lewis starts off by teaching us something about the engineering design process "... in a nutshell," as he puts it. Although the book is on the subject of creating PC-based models, Lewis cleverly uses electrical circuits and mechanical things to illustrate the different approaches ... a fact that is immensely important to those of us who have some knowledge of technology but are not yet well-versed in the mysteries of writing models.

I tried very hard to find something negative to say about *Modeling Engineering Systems* because, as my own experience for the past 25 years indi-

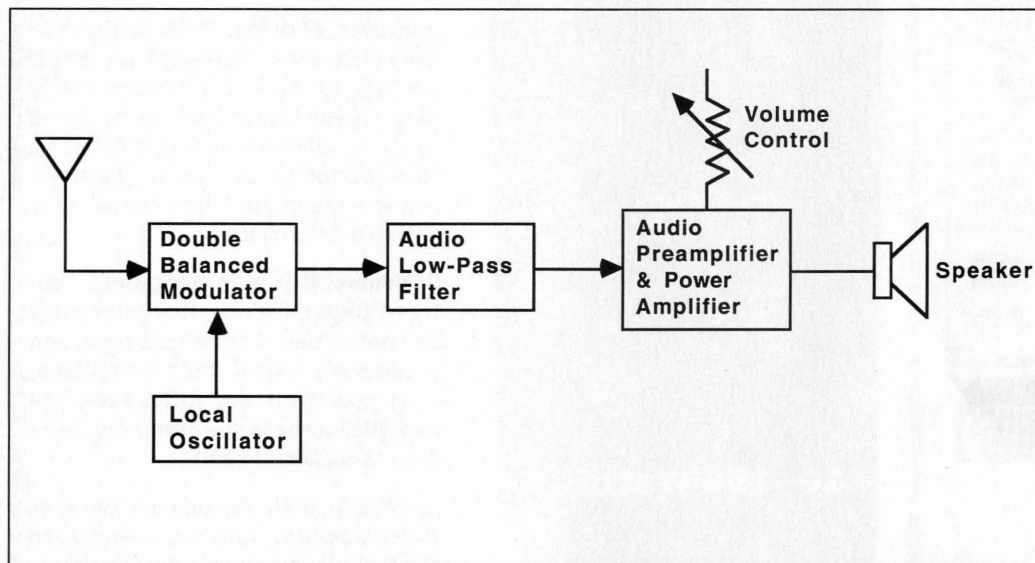


Figure 1. Block diagram of a direct conversion receiver.

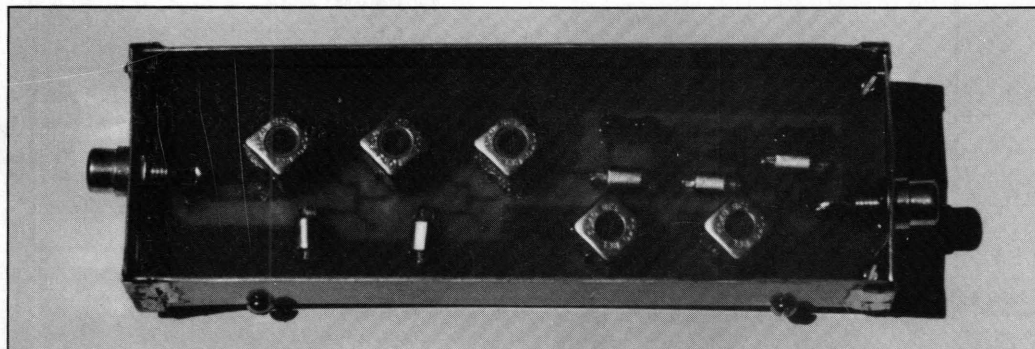


Photo A. Actual as-built filter circuit.

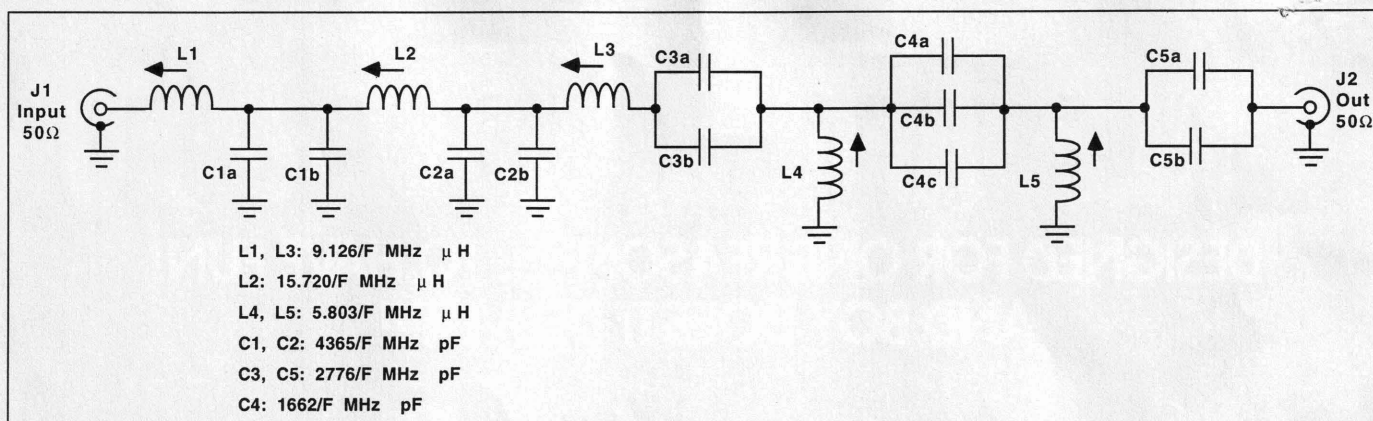


Figure 2. Bandpass filter circuit.

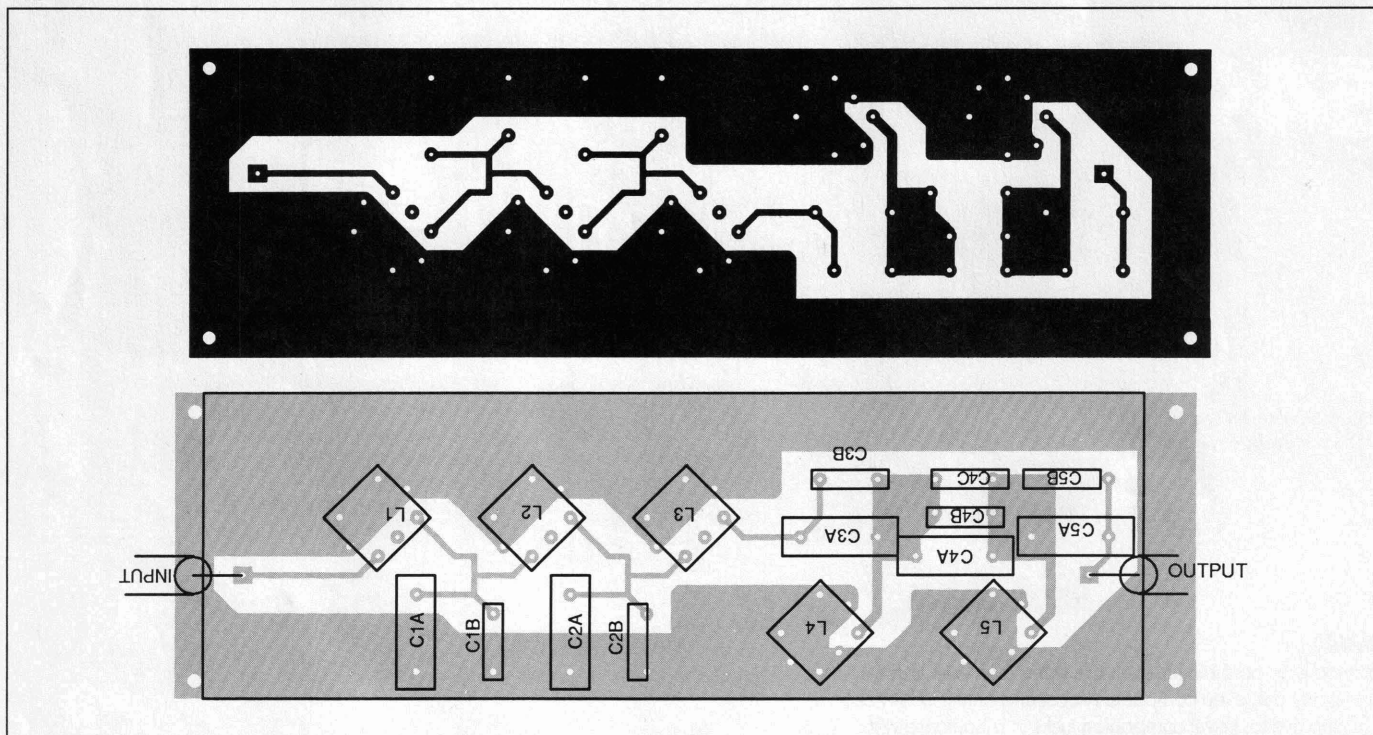


Figure 3. PCB pattern and parts overlay.

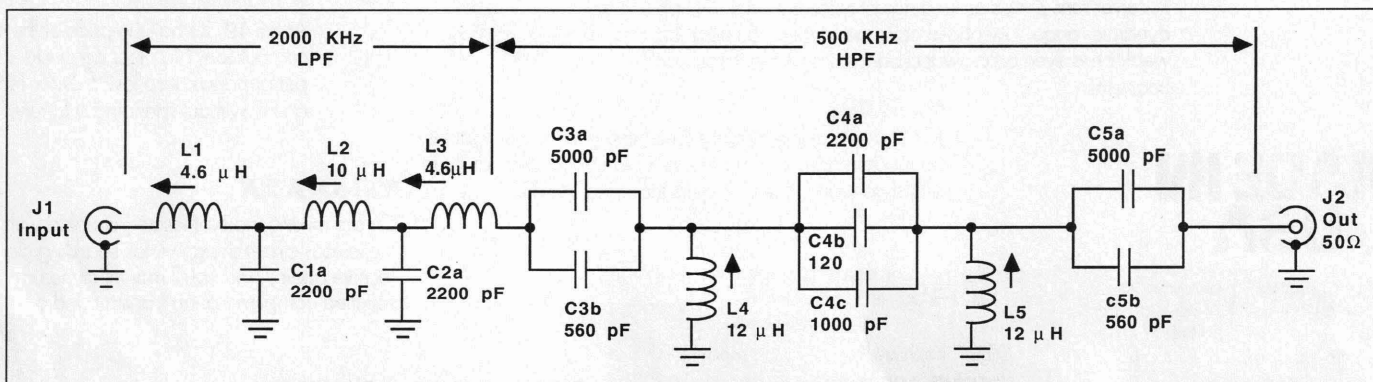


Figure 4. AM BCB version of the filter.

cates, reviewers are supposed to maintain their "tough image" by castigating the ancestry of authors, or tearing down their ego or somehow vilifying their work. But that would be the jealousy of a fellow author speaking, not my actual judgement. I really couldn't find anything missing or

wrong in Lewis' book, try as I might.

I can predict one thing about Modeling Engineering Systems, however: Forward-looking engineering and electronic technology schools will adopt Lewis' book and recommend it to their students. Backward schools, where the professors are less like practicing

engineers and more like sedentary slugs teaching running, will take note of three facts: 1) it is low-priced (\$19.95 not \$99.95); 2) it is an easy read; and 3) it is a large paperback rather than clothbound . . . and thereby reject it. After all, they will unreasonably, why give students an easy-to-

read book, that makes a difficult subject damn near easy, and that costs less than dinner out?

Seriously, though, if you have any interest in computer modeling of circuits (or any engineering system), then give Jack Lewis' book, *Modeling Engineering Systems*, a close read. **73**

ABOVE & BEYOND

VHF and Above Operation

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Frequency reference oscillator stability tips

This month I want to cover a circuit that can be incorporated into your station's test equipment for either improved timebase accuracy or as a marker harmonic generator for general testing. It can also be used in conjunction with frequency counters which have external timebase inputs to improve operation. If the frequency counter does not have an external input for the timebase oscillator, the circuit can be used to measure and correct the internal oscillator adjustment to proper operation.

With the construction of a simple TTL frequency divider, other marker frequencies can be obtained, depending on the innovation of the constructor and his needs. Stability is an important issue, and there are a few things you can do to either assemble a reasonably good unit or improve the operation of an existing one.

First, let's get into what makes a unit unstable, or the flip side for one that moves in frequency and does not stay put. The simplest circuit that can be constructed is a basic crystal oscillator. It is composed of a transistor or TTL gates and a few resistors and capacitors, along with a crystal. This circuit is usually constructed on top of a circuit board and powered by a battery or low-voltage power supply.

There are several problems with this type of circuit and several benefits. On the plus side, it's small, functions reasonably well, and oscillates

somewhere near the crystal-marked frequency. On the negative side, as it is a simple basic circuit, it will drift in frequency about the marked one due to voltage, temperature, and the loading effects on the oscillator (depending on what you connect the device to). Loading can be looked at as the same as using some of the oscillator's power in the output. When you affect the oscillator in any way, the circuit changes ever so slightly and the frequency of the crystal becomes affected by the power being used in the load. Multi-stage oscillators reduce this loading effect by drawing very lightly on the oscillator and then amplifying the signal.

There are several other considerations to be taken into account, including capacitive loading and leakage from the main oscillator to the output without going through the intended circuit. This leakage occurs when oscillator energy jumps over circuitry through stray coupling paths between the actual oscillator and the third stage of such a circuit. In this jumping over there is coupling between the first stage and the third, with minimal going through the second stage.

It would be like having 1 MHz energy being coupled into the output of a divide-by-10 circuit. In this case, there would be both 100 kHz and 1 MHz present on the output. You can see what I am getting at: signals going around circuitry and re-combining in the output.

How, then, do you design a good oscillator, and what are the

properties that make an oscillator great? Looking at some very high accuracy frequency standards, we might not be able to duplicate their excellent construction and circuitry details, but we can adapt several of these features into a good facsimile.

Attributes of a good frequency standard include very tight voltage regulation and low AC ripple (hum and noise) on the DC component used to power the oscillator. A stable temperature environment for the crystal oscillator and associated oven temperature control circuits are also needed. To improve temperature stability, there are usually two ovens. One is a slow time-constant oven directly affecting the crystal oscillator. The second

heater is a faster time-constant oven wound over the first one to help stabilize the environment of the assembly.

The inner oven might have a temperature cycle that takes a day of operation before it is able to hold a temperature to a very small range. The outer oven might only take a few hours to achieve this same or nearly same temperature stability. All this is going on to hold the crystal to a single exact temperature and prevent it from being affected by outside effects such as the garage door opening or drafts.

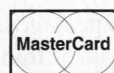
To protect this assembly from drafts and other small temperature effects, the inner and outer

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oven are inserted into a controlled-temperature chamber. In many circuits, the device used is nothing more than a sophisticated custom-made Thermos™ bottle with a hole through the sealing cap for bringing leads out to run the circuitry.

Actual construction of the device containing the oscillator would look like a three dimensional sandwich containing the inner oscillator circuit, protected with some insulation, and covered by the first inner oven. Then more insulation such as Styrofoam™ and the outer oven and still more insulation like a thermal blanket to hold the unit snugly in the Thermos bottle. It's kind of like constructing a ship in a bottle, but making sure that everything is insulated (thermally) from each other, and that all voids are minimal.

Can we duplicate the circuitry or is this too complex? Yes we can, if we start out with an oscillator circuit that is not a simple unit but has several stages to isolate the oscillator from the output, and incorporates voltage regulation! We ran into a suitable unit, a 10 MHz TCXO oscillator that had internal

temperature adjustment and 10-volt voltage regulation. The oscillator was made to function on +12 volts and ran on 35 mA of current so as to not produce much heat on its own account. There is a variable capacitor access hole on the oscillator side to set the unit to frequency.

Construction can be as simple as placing an insulated unit on its own in a Thermos bottle and letting the ambient temperature do its trick. Or you could create a very slow time-constant single oven to maintain some preset temperature above a normal high ambient, say 90 degrees. Why slow time-constant temperature control? Because a fast temperature control will overshoot its mark and be running continually, either trying to catch up, or (when it cools down), trying to reheat the unit.

A very slow control reacts over a longer time period and will ultimately remain quite fixed at some mean temperature point you select. This will continue when inserted in a Thermos bottle, as it, and all the components in the bottle, will maintain the same temperature.

If you have an oscillator already, or have put together something similar to this, I

outlined previously how you ensure you have a good accurate unit.

There are several methods to accomplish the necessary calibrations for achieving a specific point of stability in your oscillator. Time and frequency standards are broadcast by the government's National Bureau of Standards, now renamed NIST. The calibration of your oscillator can be accomplished by comparing it with one of these transmissions. The simplest of them all is WWV, which broadcasts frequency standard information at 5, 10, 15, and 20 MHz exactly. Their frequency is compared at the source to extremely accurate reference oscillators.

To compare your oscillator, you only need a receiver capable of receiving these frequencies and to make an audible beat note adjustment minimizing the difference between your oscillator and WWV. This method is capable of making an adjustment that is accurate to 1 part per MHz. Not too bad, but this is just the start of what could be a very accurate comparison.

The reason you are not able to make exact reference is that on these frequencies transmission via the atmosphere affects the exact frequency and changes it slightly. A more accurate transmission is at the VLF or

Very Low Frequency of 60 kHz. Here the atmosphere does not impart as much of a change on the signal and better resolution can be obtained. Other sources include LORAN, GPS (Globally Positioned Satellite), or even calling up NIST on the telephone and obtaining voice time readings. There is even a service on the Internet that will adjust the clock of your computer to exact or nearly exact time.

More information can be obtained from the NIST information published on their Internet page at [<http://rio.bldrdoc.gov/timefreq/>]. This is the best informational page covering most of the frequency standard information that is available—and it's free. Can't beat that for good information, so give it a look-see.

Just in case some of you can't locate a suitable oscillator for trying some of these methods that I have covered, I will be glad to provide one for experimentation. I have come across a suitable oscillator with specs that are quite good. I have a large number of "pulls" that were rejected for one reason or another. I suspect these units fail to meet original temperature specs over a great range. I have short-term tested them and junked any that fail to function, are way off frequency, or have any other malady. Instead of

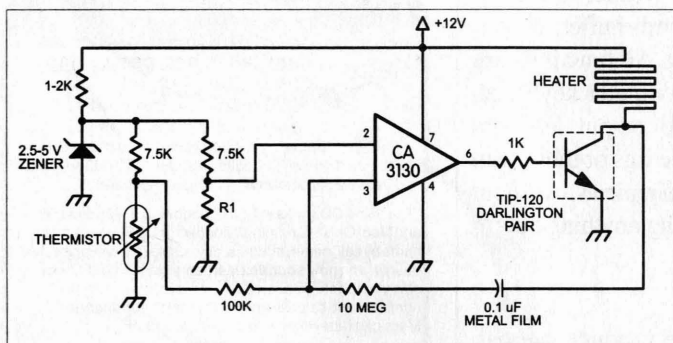


Fig. 1. Schematic diagram of simple one-stage temperature control circuit using a 3130 op amp. R1 is selected to reflect the temperature of the thermistor's resistance at the calibration point selected. The heater can be a large number of turns of #36 enamel wire or nichrome resistance wire to provide a suitable load (10 to 20 Ω) for the TIP-120 Darlington transistor. The thermistor selected was a 10 k Unicurve™ device, but others will work. It should be in direct thermal contact with the oscillator's metal container for best results. R1 = thermistor temperature at regulation point. For example, if thermistor used is 678 Ω @ 100°, then R1 = 678 Ω .

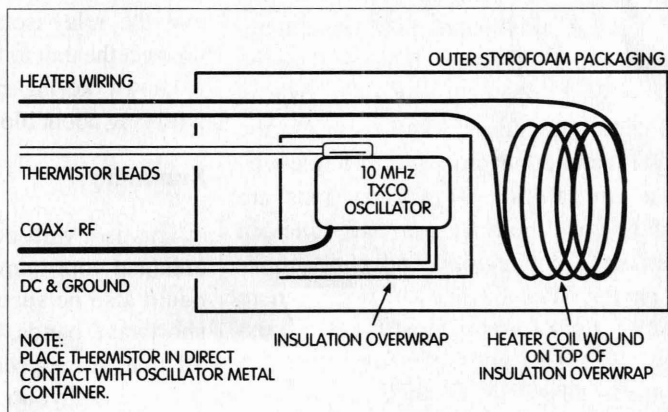


Fig. 2. Diagram of TXCO 10 MHz oscillator with oven wound on outer case of main oscillator. Wire insulated from main oscillator and complete unit packed in Styrofoam insulation before insertion into temperature-controlled chamber (Thermos). Cut small holes for leads to exit bottle through plastic cap.

let you tailor it to your specific use and style, just like a typical TNC. Although it doesn't use DSP technology, the reader employs a CMOS digital phase-locked loop and analog filters controlled by the on-board micro. For instance, you can set the noise threshold and the analog gain, and even the PLL bandwidth has wide and narrow settings! That sort of close interaction between the analog, PLL, and micro-processor stages is what gives this device such an advantage over the modem-and-software approach.

Upon first opening the very well done manual, all those options looked a bit daunting. As it turned out, though, the thing worked quite well with its default settings, and I didn't have to set anything beyond my comm port in order to view code on my screen.

The device can receive at speeds beyond 35 wpm. It has a 512-byte (character) transmit buffer, and about 8 k of saved message buffers. It can transmit files from disk (so you won't have to manually send that brag file over and over), and will even automatically ID your station every 10 minutes. It can log received data to disk. There's even an on-screen tuning indicator, along with the tuning LEDs on the unit. As a keyer, it offers relay-isolated contacts, and will send at whatever speed is being received unless you tell it otherwise.

As I mentioned, the manual is very well done. It's in 8-1/2 by 11-inch format and stapled together, which seemed chintzy to me until I started using it. Then, I found it handy to have such big pages, and the lack of a binding let the book stay open to whatever page I picked. Excellent drawings of the board and connector layouts made wiring up cables easy. Connections are shown for both 25-pin and nine-pin PC-style connectors. Sadly, no hook-ups for Mac are shown, nor is there any Mac software. Most hams are using PCs, so I guess they concentrated on that platform. Explicit instructions are given for making all of the connections. I, for one, regard good documentation as a tremendous plus for any product, and this one's a winner, despite a few typos.

Let's do it

OK, I had everything wired and running. I powered up the board from my benchtop 12-volt supply. Now, I tuned the band and, hey, it's reading code! Signals my old modem-and-software system wouldn't have made any sense out of, this thing decoded very decently. The LEDs made it easy to tune the rig to the optimum point for the device's filters, even without glancing at the on-screen tuning indicator.

Performance

No code reader can beat the human brain, nor should you expect it to. Well-sent code was displayed nicely, but even the best code often had words run together. Weak signals decoded better than I'd expected, but there were plenty of errors when signals faded down low or QRM got heavy. The unit did a surprisingly good job of reading hand-sent, straight-key code, but it did not do well with "swung" Morse, in which the sender alters the normal 3:1 timing relationship of dots and dashes. In all fairness, my own personal "wetware" decoder doesn't do too well with that either! Luckily, few operators swing their timing anymore (thank goodness).

The keyer portion of the device is much like other memory keyers, which is to say it's quite nice. It lets you put received data, such as name, QTH, or callsign into the transmit buffer, so you won't have to retype it. All functions are controlled from the computer keyboard, and the relay-isolated output lets you connect the unit to tube rigs or those with unknown keying requirements, without worrying about blowing anything up.

Summary

This is a well-done product that can enhance your ham radio experience. It could also be successfully used on the shortwave bands, letting you see what all that high-speed code is all about. (Of course, you're out of luck if the signal is encrypted.) It does a very respectable job of decoding CW, and it beats the old modem-and-software routine any day of the week. If you like CW, check this thing out!

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is constructed out of several layers of spring material and backed up with a heavy brass or plated-steel armature for the switch contact rotary section. This type of switch is typical of a surplus heavy-duty RF switch removed from older military surplus. Usually it is so heavy it must be mounted on a panel in order to operate the manual switch from one position to the other. It switches into position with a loud "snap" sound, making very evident it switched.

I would hesitate to use this type of switch above 20 MHz and rate it iffy at best near that frequency. An improvement would be to mount the rotary switch in a metal box and fix connectors for each switch contact. As you look inside the metal box of such a switch and its wiring, you can see that there is a certain amount of coupling between each section of the switch. In other words, there is still some coupling (poor isolation) between sections of the switch that are in proximity to each other. Think of the operation as coupling between a tangle of clip leads scattered on your workbench.

To test this isolation, you just run a test signal through the closed contact path and test at the open contact to see how much of the signal leaks through to here and its connecting leads (see **Fig. 2**). What you want to see is lots of isolation and very little signal leaking through. The measure of the loss is the rating of isolation, usually expressed as dB isolation. In this simple switch loss is low, giving it a poor rating above 20 MHz because of poor isolation as frequency is increased above a point. Operation at lower frequencies is quite proper.

The reason you want high loss between the switch contacts is that a transmit path might have 100 watts of power going to the antenna. The open contact is the receiver, and if the isolation is poor, an appreciable amount of power will be coupled into the front end of the receiver. This is not what we want to happen. With a receiver sensitive to minus 100 dB and transmitters with output powers in the 100 watt category (+50 dB), it is safe to assume that

a great switch would have isolation in the 50 dB or better range to protect the equipment (receiver) it is switching.

Coupling (poor isolation) is the bane of any coaxial relay circuit. If the frequency is low, the extraneous wiring can be tolerated because the wavelength at these lower frequencies is quite lengthy and short hookup wiring techniques are tolerated. As the frequency rises, shorter wiring methods must be used to limit excess wiring and its associated coupling. All is not lost, as there are switch layouts that help to minimize coupling between switch elements and make them virtually invisible to each other.

What layout can we use to provide low coupling or high loss between adjacent elements of the switch? Well, when we think it out, it can be shown that the best switch would duplicate as best it could a manual coaxial connection. The switch would look in this scenario like a manual coaxial switch panel. The switch panel is one that is entirely made up of coaxial connectors and one patch cord (coaxial). When you wanted to select a new port, you would have to unscrew the connector and move the cord to the new port connector. Not very practical, but very efficient in minimizing coupling and effecting very high isolation, which is excellent.

What is needed is a mechanical contrivance that duplicates this action to obtain the very best in isolation and at the same time maintain almost zero coupling between ports—just like the previous coaxial connector-and-cord scenario. I am not dreaming, as some of you might suspect by now, but rather I am just trying to make you aware of what is going on and how important isolation can be.

The switch that conforms to this design principle is made by Transco. Its operation is exactly what was previously described in a manual-patch cord scenario. The switch I tested has manual operation and six possible output ports. It exhibits all the quality of the coaxial connector-and-cord operation in a very compact rotary coaxial switch. It uses a spring contactor coaxial hairpin,

HOMING IN

Radio Direction Finding

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Low-cost monitor for the NorthScope

"Is that radar in your car?" I'm not surprised when visitors to the starting point of our hidden trans-

mitter hunts (T-hunts) ask that question. The lingering yellow glow of the lines on my cathode-ray tube (CRT) display are reminiscent of radar scopes in pre-computer times. But these lines are signal bearings, not aircraft tracks.

Continued

allowing contact in a coaxial environment with only one contact between the selected main input/output connector. To switch it to another position by manual rotary action, it has a cam action to unseat connections before reseating into the new selected position.

There is virtually no coupling between the main and unselected switch ports, due to the excellent shielding between master and selected rotary contact all done in a coaxial environment. This minimum coupling is accomplished by this cam-operated coaxial hairpin internal to the switch. In all respects this hairpin is just like unscrewing a connector and transferring it to another connector.

As far as isolation goes, the other connectors might as well be in the junk box as you cannot see them with this type of mechanical contact switching. The other unused connectors are out of the picture, electrically speaking. The selected path and its connector are totally coaxial and shielded from everybody else.

The Transco switch is quite robust; I have used it to 450 MHz with no problem. This switch must be bolted to something sturdy, as it does require a few pounds of pull to rotate the switch. You need this hard mounting in order to turn the switch by hand. It has quite a stiff cam action and produces a sharp "snap" sound (of the kind mentioned earlier) when the coaxial hairpin is reseated.

The relay that I show in **Fig. 1** and use at 10 GHz is quite small and uses miniature SMA coaxial connectors. The coax cable that is used has Teflon™ insulation and relatively low loss at these frequencies. As you can see in **Fig. 1**, when the relay has current flowing through the coil it attracts the armature to the pole piece on the coil. This activates the armature in a teeter-totter type of function and uses insulated push rods that raise one end and lower the other end of the switch contacts.

The unused switch contact is pulled toward the top of the switch's enclosed chamber and grounds out on top of the switch compartment. The other element is pushed into contact with the previously open contact and the center main contact. The process reverses when relay current is removed. The internal compartment where this switching action takes place is much like a very short section of air dielectric coaxial cable with the exception that its internal dimensions are square and the impedance is 50 ohms. The impedance is determined by the ratio of inner to outer conductors.

Think of this inner-to-outer ratio as quite similar to the ratio between coaxial cable and its inner and outer conductor. The action is quite the same. When the switch duplicates as closely as it can the coaxial environment and the internal elements are a fraction of a wavelength at the frequency of interest, it will function well.

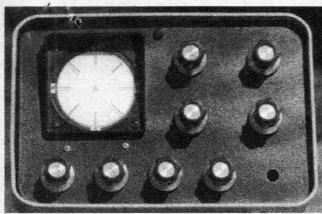


Photo A. The Heathkit Scanalyzer after repainting the front panel and replacing the grid screen under the CRT bezel with a transparent compass overlay. Next comes labeling. The hole at lower right will be filled by a future control.

This month's "Homing In" is the final installment of a series on north-referenced bearing readouts for radio direction finding (RDF). July's column covered remote heading sensor technology. An inexpensive fluxgate compass module gives you a dashboard indication of your mobile antenna mast position with respect to true north. With it, you can quickly tell if the bearing to a hidden T is steady or shifting as you drive on winding roads in the dark.

"Homing In" for August described an analog multiplier circuit that combines fluxgate compass and receiver S-meter signals to produce a "north-up" display of signal strength versus direction. I call it the NorthScope. Its polar plot simplifies the task of separating the direct signal from multiple reflections (multipath) in urban and hilly areas.

Like a radar scope, a north-up display must have persistence of several seconds so that the operator can "stack up" traces. That makes it possible to tell the difference between fluttering reflections and more stable direct signals. Ordinary oscilloscopes and computer monitors are not suitable because their persistence is only a few milliseconds. Surplus waveform storage oscilloscopes and medical monitors make fine readouts, but they require 110-volt AC power and are somewhat expensive.

A recycled boat anchor

I was searching the swap meets for parts to build a home-brew display when I discovered a '60s-vintage Heathkit SB-620 Scanalyzer. There, in one case, were

all the hard-to-find items I needed—including the long-persistence CRT, mounting hardware, mu-metal magnetic shield, socket, panel, bezel, high voltage supply, and beam controls. What's more, the cabinet had plenty of room for the interface circuit with its front-panel controls.

Constructing a CRT display from scratch is possible. I built a bigger one for my high school science fair years ago, but surplus scope parts were much easier to find then. Nowadays, the modified Scanalyzer approach is far simpler and cheaper. They show up regularly in estate sales and at the three computer/electronics swap meets in my area. A recent issue of *Amateur Radio Trader* magazine had an SB-620 listed at \$75. This appears to be a typical "street price." Be sure to get the manual, too.

The Scanalyzer need not be in full working condition, but the high voltage supply, CRT and beam controls (INTENSITY, FOCUS and ASTIGMATISM) should be functional before you begin the modifications. There are Internet mailing lists and Web pages devoted to "boat anchor" equipment and Heathkits that can help you locate an SB-620 and repair it if necessary. You will find links to these resources at the "Homing In" Web site. My site also has information on CRT sources and high voltage supply schematics for those who choose to build a monitor from the ground up.

The SB-620 has a 3RP7 high-persistence CRT with a 2-5/8-inch-diameter face (**Photo A**). This is quite suitable for a mobile display. I had considered the next size larger CRTs for the project, but electrostatically focused five-inch tubes are about 17 inches long, necessitating a deep enclosure. They also require much higher acceleration and deflection voltages than the 3RP7.

WARNING: Use extreme caution when working on your monitor. Voltages high enough to cause serious injury or death are present in the power supply, beam and deflection circuits. There is no substitute for caution, prudence and clear thinking when working with high voltage. Mount all circuits inside a grounded metal enclosure so that high voltage points are never exposed to the user during normal operation.

Turn off all power and wait for capacitors to discharge fully before removing the cover. With power off, short all capacitor terminals in the high voltage supply to ground with the metal shaft of an insulated screwdriver before touching any circuits. Observe all safety precautions including keeping one hand in your pocket when making adjustments and measurements with the cover removed. Do not work on HV circuits when fatigued or under the influence of medication or intoxicants. Measure high voltages only with instruments and probes designed for this purpose.

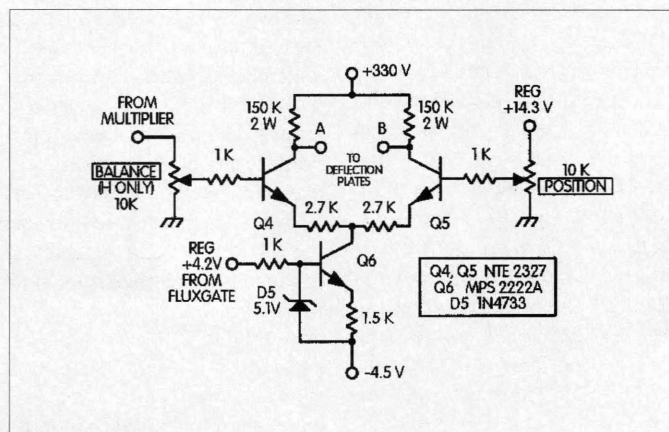


Fig. 1. Schematic of the new deflection circuits. Two are required, each with a spot positioning potentiometer on the front panel. A and B outputs of the horizontal (X) circuit go to the orange and yellow CRT socket wires, respectively. A and B outputs of the vertical (Y) circuit go to white and violet socket wires.

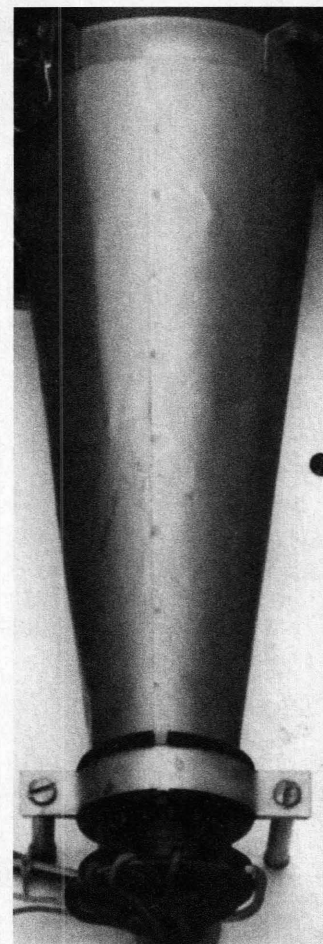


Photo B. Rear view of the CRT inside its mu-metal magnetic shield, with socket. Leave the shield on the CRT to prevent spurious deflection due to magnetic fields from the transformers.

Use care when handling a CRT due to its high vacuum. Do not

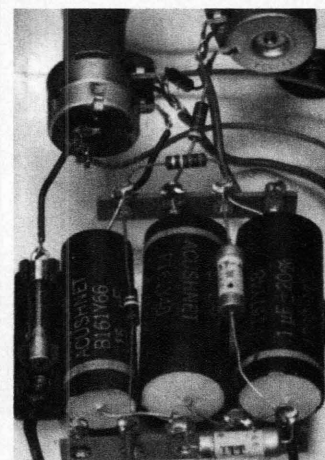


Photo C. Components in the CRT cathode supply are as originally described in the SB-620 manual. Note the long insulated bushing on the INTENSITY control. The piggyback switch is rewired for +12 volts.

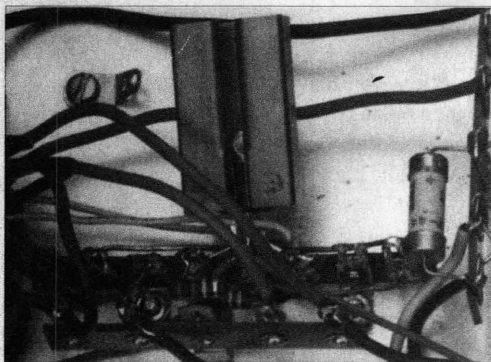


Photo D. This series regulator stabilizes the acceleration electrode and deflection supply voltages to prevent spot movement as vehicle voltage fluctuates.

strike, scratch or subject the CRT to more than moderate pressure at any time. A fracture of the glass could result in an implosion capable of causing injury.

Tubes out, ICs in

The first step in the SB-620 conversion is to remove all of the RF circuits, tubes, tube sockets and associated components, except for the CRT of course. The INTENSITY and FOCUS potentiometers are used as is. All other front panel controls are no longer needed. Replace and rewire them with switches and potentiometers for the new INPUT, BALANCE, HORIZONTAL POSITION, VERTICAL POSITION, SIZE and OFFSET controls.

Disconnect the violet, orange, white and yellow CRT socket wires at the chassis end, leaving the other socket wires connected to the supply (**Photo B**). Leave all power supply components and wiring intact for now. Note that the INTENSITY control has an extended insulated shaft and bushing. For high-voltage safety, be sure this is assembled per the construction manual and **Photo C**.

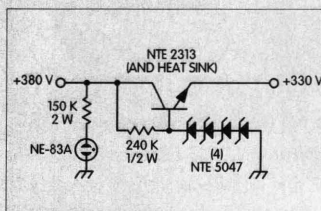


Fig. 2. Schematic of the +330-volt regulator. The NE-83A indicator and 33k resistor are on the original SB-620 front panel.

Build and install new X and Y deflection circuits per **Fig. 1**. A perfboard photo is in last month's "Homing In." The previously disconnected CRT socket deflection wires connect to these circuits. Except as noted, added resistors are quarter-watt. Higher wattage

resistors must be used in some places for their greater voltage standoff capabilities.

There is only one BALANCE control, used in the horizontal deflection amplifier. It equalizes the X and Y path gains, compensating for the slightly higher horizontal deflection plate sensitivity in the CRT. When performing the alignment described last month, adjust the BALANCE control to obtain a perfect circle trace instead of a horizontal or vertical ellipse.

Only a few changes are required to the rest of the power supply and beam circuits. Rewire the on/off switch and main fuse to open the vehicle's DC input instead of the main transformer primary. The DC fuse should be two amperes fast-blow. Negative supply (-4.5 V) for the LM324 comes from the original DC heater source (D8-D11 and C54).

Deflection supply and CRT astigmatism (Anode #2) supply voltages are regulated for mobile operation. The +330-volt regulator circuit is in **Fig. 2** and **Photo D**. I built it on an added eight-terminal strip next to the five-terminal strip at position E on the chassis. Be sure to provide heat-sinking for the transistor, but do not ground the sink. Remove the wire from the top of the astigmatism control (lug #3, farthest from the power input grommet) to the positive end of C53-C in the power supply. Connect the regulator input to the positive end of C51. The regulator output goes to lug #3 of the astigmatism control and the common point of the four

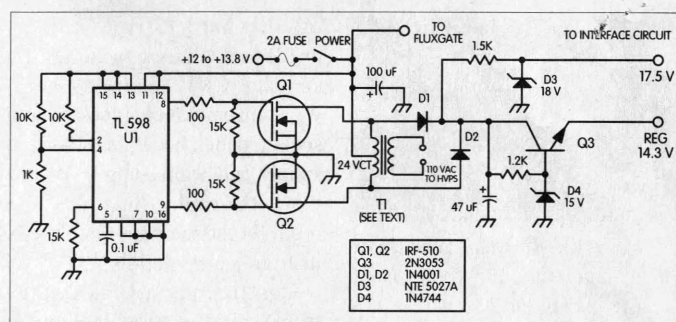


Fig. 3. Schematic of the DC-to-AC converter. Outputs are 110 VAC square-wave for the high voltage supply and positive DC for the analog multiplier circuits.

150k collector resistors in the deflection circuits.

Powering the scope from a car battery is easy with the circuit of **Fig. 3** and **Photo E**. The TL598 converter IC is available from distributors of Texas Instruments™ semiconductors. It generates alternating pulses at 500 Hz to drive the two switching transistors. Although the TL598 has provisions for voltage and current feedback, I found it to be unnecessary in this application.

Mount the transistors to the chassis with supplied insulating hardware. The transformer (Triad F-45X) is connected "backwards." Its 24-volt center-tapped secondary connects to the transistor drain pins. Output at the primary is 110 volts square wave. If you can't find the Triad transformer, you can substitute Radio Shack™ 273-1512, but output voltages will be about 5% lower.

There is no problem feeding the square-wave chopper output into the main transformer primary to operate the scope instead of a sine wave. The RMS voltage at the CRT heater is very close to the wall power value with normal +13.8-volt input. The capacitor-input high voltage supply filters produce about 35% lower voltages, compared to AC power. This is an advantage because it lowers the CRT deflection voltage requirements such that solid-state drivers of **Fig. 1** are practical. Spot brightness is more than adequate at night, but add a sunshield for daytime hunts. You can set the INTENSITY control to maximum without burning the CRT phosphors.

Except for the special components mentioned above and the NTE semiconductors, all of the

new parts should be available at your local Radio Shack. NTE transistors and diodes are sold by local parts distributors nationally. My junk box played a large role in parts selection for this project. Experienced experimenters should feel free to make appropriate substitutions.

With modifications completed and the analog multiplier circuits installed, your NorthScope should be ready to align and operate. Be sure to check your work thoroughly before the first "smoke test." Warm-up time of the CRT is about 20 seconds. DC supply current is 600 milliamperes in normal operation, not including the fluxgate compass. Follow the alignment procedures given last month for the fluxgate-to-CRT interface.

Better than a Doppler?

Fans of Doppler RDF units may argue that their method is

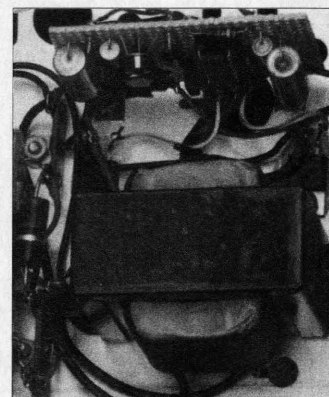


Photo E. The DC-to-AC converter circuit includes a step-up transformer, two field-effect transistors bolted to the chassis and a 2-1/2 x 1-1/2-inch perfboard with the IC and associated components.

CARR'S CORNER

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Receiving loops & loop preamplifiers: part 2

As discussed in part 1 last month, the output signal levels produced by loop antennas are typically very low. The use of a tuning capacitor can increase the levels by 10 to 200 times, but even these levels are often below what is needed. To overcome this problem loop users often provide a preamplifier to boost signal before sending it to the antenna input terminals of the receiver. Before looking at specific designs, let's take a look at the various amplifier-loop configurations.

Amplifier-loop configurations

The basic amplifier configurations are single-ended and balanced. The single-ended amplifier has only one input terminal and ground. The signal is applied between the input and ground. The balanced amplifier, which includes both push-pull and differential amplifier designs, uses two inputs. The amplifier responds to the difference between the signal applied to the two inputs.

Fig. 1 shows the most basic single-ended amplifier and loop configuration. The loops in

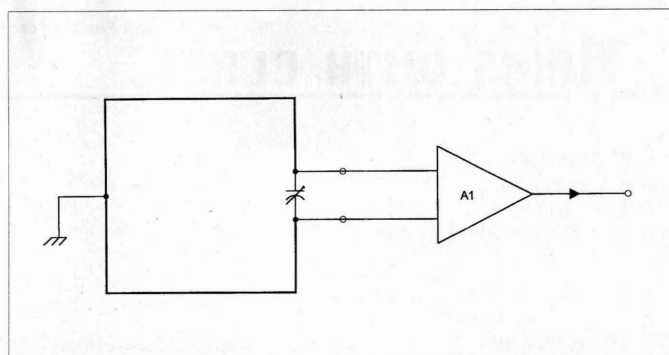


Fig. 1. Basic single-ended loop amplifier.

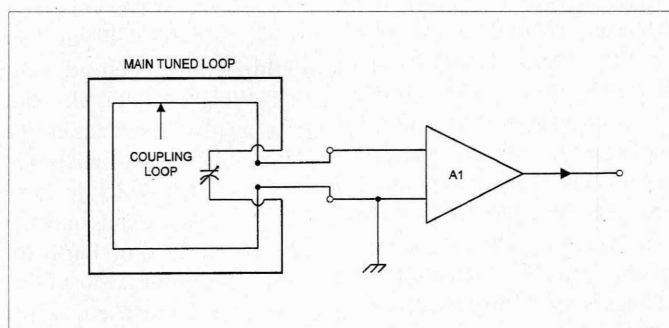


Fig. 2. Single-ended with coupling loop.

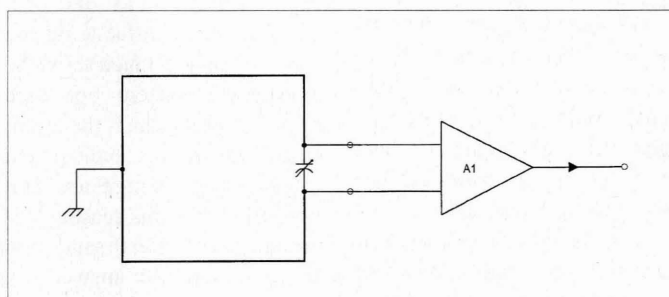


Fig. 3. Balanced loop.

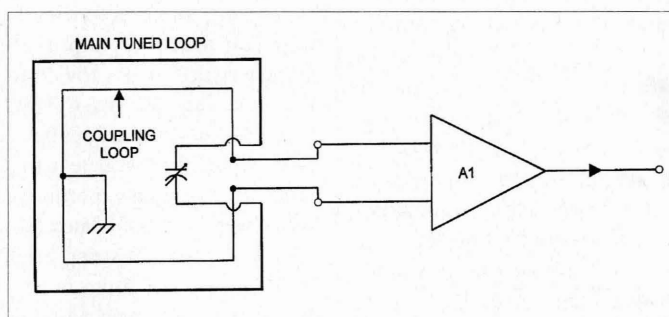


Fig. 4. Balanced with coupling loop.

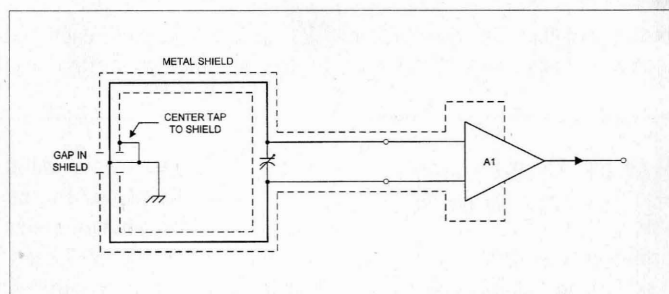


Fig. 5. Shielded.

HAMS WITH CLASS

Continued from page 53

ings), heavy white construction paper, and a bar magnet.

Procedure

Measure the length of the magnet. On the paper, draw a circle with the same diameter as the length of the magnet. This will represent Earth. Mark the top of the circle "North" and the bottom "South." Lay the bar magnet on a flat surface such as a table. Place the paper over the magnet so that the ends of the magnet are on the "North" and "South" marks. Sprinkle the filings over the paper. Gently tap the paper, allowing the filings to move. Keep tapping until the pattern stops forming. 1. How can you tell where Earth's magnetic poles are? 2. What do the lines formed by the filings represent? 3. Look in an atlas to locate Earth's geographic poles.

Are they in the same place as its magnetic poles?

A good place for an educator to get started with useful materials to use in the classroom is with CORE. This is NASA's Central Operation of Resources for Educators. CORE was established for the national and international distribution of NASA-produced educational materials in audiovisual format. The address is:

NASA CORE
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Phone: (216) 774-1051 ext. 293

Let's all remember that sharing good instructional ideas is what the "Hams With Class" column is all about. Please send me any teacher-tested things you've had success with. Include photos and we'll highlight your children having fun with ham radio.

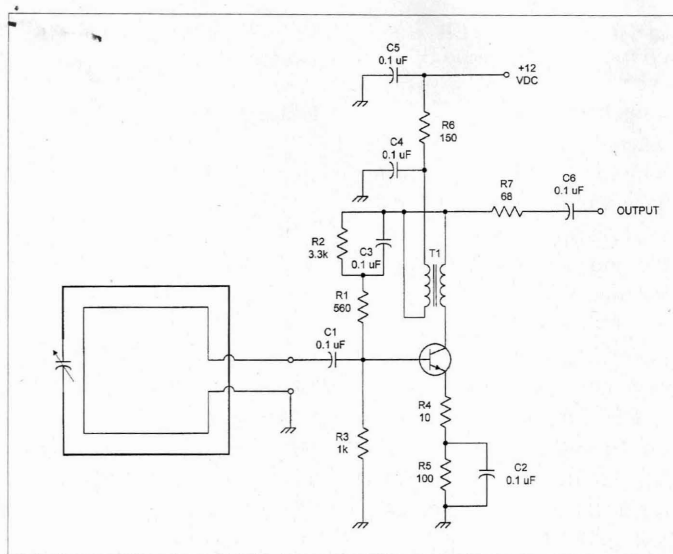


Fig. 6. Bipolar NPN transistor preamplifier.

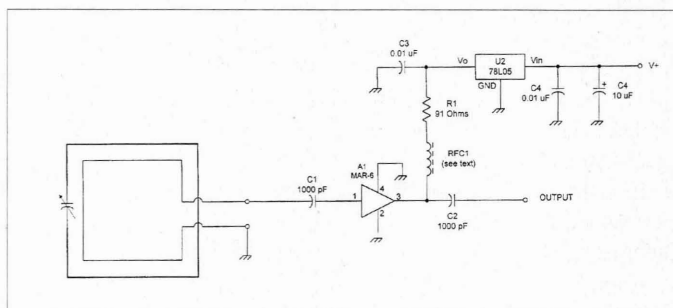


Fig. 7. MAR-6 preamplifier.

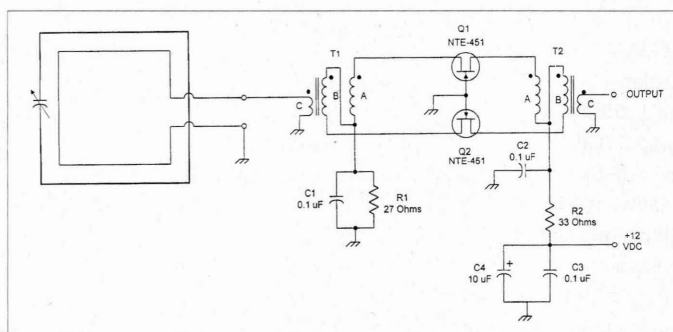


Fig. 8. Push-pull JFET preamplifier.

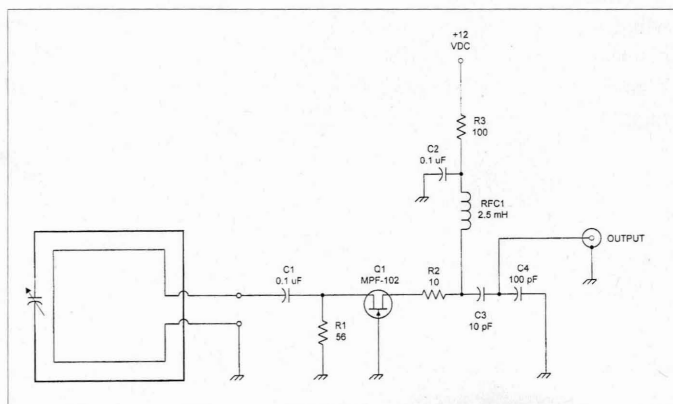


Fig. 9. Common gate JFET preamplifier.

Figs. 1-5 are shown as single-turn loops, but that is a graphical convenience, and both single-turn and multi-turn loops are intended. The loop is tuned by a capacitor. One end of the loop and the loop winding is grounded, as is one terminal of the amplifier input. A variant circuit is shown in Fig. 2. The version in Fig. 1 uses the main loop to provide signal directly to amplifier A1, while the variant version of Fig. 2 uses a coupling loop (magnetically coupled to the main loop) to provide signal to the amplifier. Both are single-ended, but in one case the main loop is used and in the other a coupling loop is used.

A balanced loop is shown in Fig. 3. This type of loop is center-tapped and the tap is grounded. This provides genuine balanced input for the differential preamplifier, A1. The same idea using a coupling loop is shown in Fig. 4.

Shielded loops are often used to eliminate pattern distortion due to capacitive coupling between the loop and its environment, and interference to reception due to local electrical fields. In Fig. 5, the loop, the tuning capacitor, and the amplifier input circuitry are Faraday-shielded by a metal barrier. A gap in the shield allows the loop to respond to the magnetic field component of the radio signals. If a center-tapped loop is used, then the center tap will be grounded either at the gap or close to the output (provided that it is electrically at the center of the loop). Tapping the loop at the output side is more convenient on multi-turn loops, while on single-turn loops the gap method is used.

Single-ended amplifiers

Fig. 6 shows the circuit for a simple preamplifier based on an NPN transistor connected in the common emitter configuration. The type 2N5719 is most often used in this circuit. Two negative feedback methods are employed in this circuit. One is an unbypassed

(R4), and the other is transformer T1. The negative feedback serves the purpose of reducing distortion and smoothing the frequency response characteristic. This amplifier is normally used from 3 to 30 MHz, but with suitable selection of transformer and capacitors a different range can be accommodated.

The unique thing about this circuit is that it has a 50 ohm input impedance and a 50 ohm output impedance, and so will interface nicely with the coupling loop and the antenna input of the usual receiver antenna input.

Another single-ended amplifier is shown in Fig. 7. This circuit is based on the Mini-Circuits MAR-6™ integrated circuit amplifier. It is a low-noise amplifier that provides about 20 dB of gain over the frequency range of from near-DC to 1,000 MHz. Capacitors C1 and C2 are used to couple signal into and out of the amplifier, while blocking DC.

The DC power is applied to the MAR-6 integrated circuit through the output terminal. The DC path includes R1, which limits current, and RFC1, which peaks the gain at high frequencies. The DC source voltage is supplied through a low-power 78L05 100-mA, five-volt, three-terminal voltage regulator.

The RF choke (RFC1) is used as a peaking coil, and is optional in some applications. The usual practice is to provide a 1 uH inductor for this purpose. In higher frequency amplifiers (VHF and above), the RF choke might be replaced by a ferrite bead slipped over a short piece of #22 or #24 hook-up wire.

The amplifier in Fig. 8 uses a push-pull circuit to improve dynamic range and reduce distortion, but from the loop's point of view it is single-ended. The single-endedness is provided by the input transformer, T1. Windings "A" and "B" are used to drive the emitters of the JFET transistors (Q1 and Q2). Winding "C" is used to couple to the

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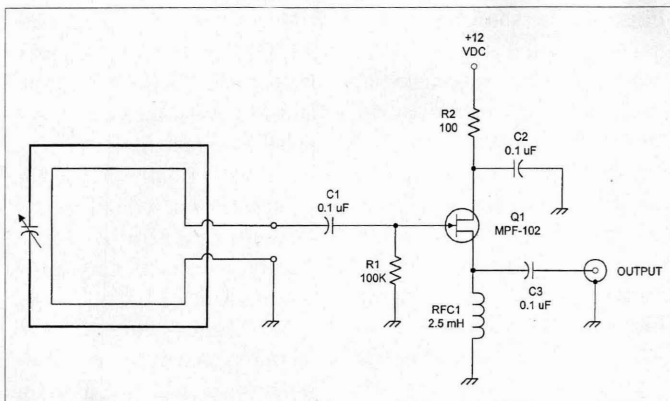


Fig. 10. Common source JFET preamplifier.

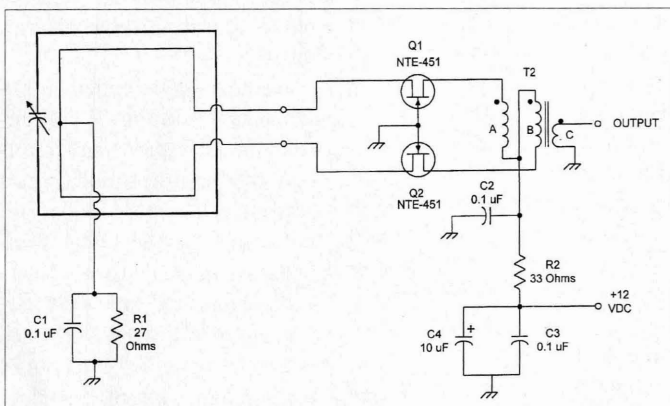


Fig. 11. Loop integrated push-pull preamplifier.

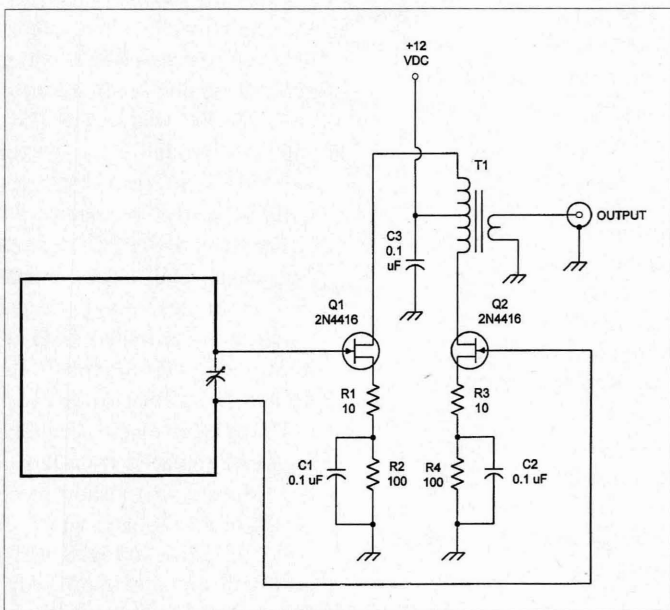


Fig. 12. Differential JFET preamplifier.

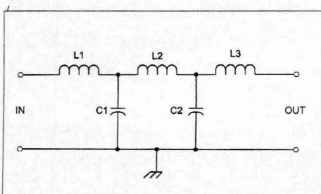


Fig. 13. Low-pass filter.

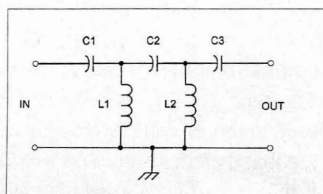


Fig. 14. High-pass filter.

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loop. If the coupling loop of the antenna is used (as shown), then winding "C" is a low-impedance winding. If, on the other hand, the main loop is used, then the impedance of winding "C" must be higher. Windings "A" and "B" must match approximately 1,000 ohms, while T2 must match 1,700 ohms.

The transistors used for the circuit of Fig. 8 are either 2N4416 devices, or the service replacement equivalents NTE-451 and ECG-451.

A slightly different design approach is used in Fig. 9. The amplifier device is an MPF-102 JFET device connected in a common gate configuration. Signal is applied across the source-gate terminals, while output signal is taken across the drain-gate terminals. Because this circuit is inherently low impedance, it is used with the coupling loop approach to antenna design.

Capacitor C1 is used to block DC bias, provided by the voltage drop across R1, from being shorted to ground through the loop. The output circuit consists of a capacitor voltage divider, C3/C4. This circuit matches the relatively high impedance of the Q1 drain to the 50 ohm antenna input used by most receivers.

Another JFET preamplifier is shown in Fig. 10. This circuit uses the same device (MPF-102) as the previous circuit, but in the common drain configuration. The signal is applied across the gate-drain terminals, while the output signal is taken from the drain-source terminals. The source circuit, which is used as the output, has a series-connected RF choke. This part keeps the source terminal at a high impedance for RF,

while allowing it to complete the circuit for DC.

Balanced amplifiers

An example of a balanced amplifier is shown in Fig. 11. This circuit is a variant on the theme of the Fig. 8 circuit in which the input transformer (T1 in Fig. 8) is replaced by the loop. Because the circuit is a common gate push-pull amplifier, the coupling loop must be center-tapped in order to provide balanced (but out of phase) input signals to Q1/Q2. As with the previous version, the DC is applied to the drain terminals of Q1 and Q2 through the windings of output transformer T2.

A common-source differential amplifier is shown in Fig. 12. The active devices, Q1 and Q2, are 2N4416 JFET transistors. Source bias is provided by resistors R1/R2 for Q1 and R3/R4 for Q2. A small amount of negative (degenerative) feedback is provided by unbypassed source resistors R1 and R3.

The output side of the circuit uses a push-pull transformer (T1). Such transformers are center-tapped. The DC power is applied to the center tap, and then routed to Q1 and Q2 through the windings of the transformer. Because of the center-tapped loop, and center-tapped output transformer, Q1 and Q2 operate out of phase with each other.

Input filtering

One of the more difficult to solve problems in radio receivers is overload from out of band signals. The frequencies that are most often used with loop antennas are often subject to strong overload from AM broadcast band stations, local amateur radio stations, and other transmitters. These signals can cause problems in the receiver even if they are not heard in the output, and thus need to be eliminated. When a preamplifier is used between the receiver and antenna, the problem is only made worse.

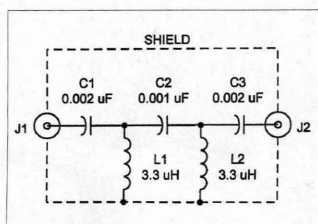


Fig. 15. Compromise.

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Mobile, Portable and Emergency Operation

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Sometimes people wonder if your ideas and suggestions are noticed. Trust me, your thoughts are the best inspiration. For example, I received an E-mail message from Walter K5KNE,

who offered some real food for thought about emergency communications. He pointed out that many hams are interested in being of service in an emergency, but are not affiliated with any

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The usual (and most effective) method for eliminating these problems is to put a filter either ahead of the preamplifier or between the preamplifier and receiver antenna terminals. Even if no preamplifier is used, if a problem is experienced then the filter can be used between the antenna and the receiver.

Figs. 13 and 14 show the low-pass and high-pass filter circuits, respectively. The low-pass filter passes frequencies below its -3 dB cut-off frequency. Such filters might be used on an LF or VLF receiver to prevent overload from AM BCB stations. In Europe, where both the medium wave (0.520–1.710 MHz) and LF band (145–280 kHz) are used for AM broadcasting, the problem is especially severe. Such cases may require two filters.

These filters are designed for 50-ohm input and output impedances, which is the standard impedance for most RF systems. The values for the components in **Figs. 13 and 14** are found by dividing the following constants by the desired cut-off frequency in megahertz (MHz). For the low-pass filter (**Fig. 13**): L1

constant = 5.6; L2 = 10.6; L3 = 5.6; C1 = 3300; C2 = 3300. For the high-pass filter (**Fig. 14**): L1 constant = 5.36; L2 = 5.36; C1 = 3900; C2 = 1800; C3 = 3900.

For example, suppose we want to design a high-pass filter to prevent AM BCB signals from overloading a medium wave shortwave receiver (3 to 7 MHz). We might select a -3 dB cut-off frequency of, say, 2 MHz. The component values for the high-pass filter then would be:

$$\begin{aligned} C1 &= 3,900/2 = 1,950 \text{ pF} \\ C2 &= 1,800/2 = 900 \text{ pF} \\ C3 &= 3,900/2 = 1,950 \text{ pF} \\ L1 &= 5.36/2 = 2.68 \text{ }\mu\text{H} \\ L2 &= 5.36/2 = 2.68 \text{ }\mu\text{H} \end{aligned}$$

A slight variation on the theme is shown in **Fig. 15**. This circuit is a "practical" compromise that allows the use of standard value components. The capacitors are 0.001 μF and 0.002 μF units, both of which are standards. The 0.002 μF capacitor can be made by paralleling two 0.001 μF units, if desired. The 3.3 μH values for the coils are a standard value. Although they can be made from toroid core and a bit of wire, these coils can also be purchased from Toko™ dealers and other sources. 

public service or other emergency agency. They expect to show up when an emergency occurs and be of assistance. As Walter put it, "The average well-intentioned ham doesn't realize that he is just another sightseer with a radio—unless he has a 'real job' with a responding agency." I've been thinking about his comment for the past few weeks as much of the nation has been facing *El Niño* weather problems. In my area SkyWarn has been activated a number of times as rain and high winds have led to flooding and several tornadoes.

Our SkyWarn system works pretty well, with one ham acting as net control, while another works with the weather bureau, and the majority act as spotters spread throughout the county. As the meteorologists at the National Weather Service become aware of developments or see a suspicious radar image, they may ask if anyone in that area can actually see and accurately describe the conditions. The SkyWarn weather spotters are one more data point the weather experts use to make a forecast. If a watch (conditions which are conducive to developing into a particular type of storm) or a warning (the storm, such as a tornado, has actually been observed) is declared, that information is then released to the public through local radio and television stations.

In many cases, though, well-meaning hams *do* act as sightseers with radios. There are the inevitable fair weather reports, or a report to the net that a television station is broadcasting a particular forecast. As amateurs we know that a signal can be lost if the noise level rises; much of what is being transmitted is noise and nothing more. What this says to me is that these folks desperately want to be involved in order to help out, but they don't have a specific function assigned. In their attempt to help out they are at significant risk of making matters worse. Imagine if our firefighters or police

officers responded with the proper equipment, the best of intentions but no clear understanding of precisely what they needed to do or to whom they would report. As the old saying goes, "If you're not part of the solution, you're part of the problem."

The key is to know exactly what you're expected to do, for whom, well in advance of the emergency situation. Area coordinators for disaster services need to prepare with this perspective rather than assume that people will be identified, located, assigned and trained in the moments after a disaster hits. Individual hams need to find out what roles they can expect to play. Since there are so many, a well-informed ham can review the possible assignments, and then apply or lobby for one of those which looks interesting.

How do you determine what is available? Obviously, experience is the best teacher, or you can talk with someone who has been involved in emergency communications for a number of incidents. To get a more global perspective you can turn to a number of books addressing disaster communications which have been published in the past few years.

Among those that are general in nature, *Guide to Emergency Survival Communications*, by Dave Ingram (Universal Electronics), is a fun book to read. It is aimed at those who want to "receive accurate and timely information in time of crisis." These are non-hams and potential hams, but the book reminds us of how many people will be listening in on our efforts to support an emergency. The preface details how Murphy's law is paramount in disaster communications.

When the Big One Hits ..., by Jerry Boyd KG6LF and Jay Boyd KN6BP (World Radio Books) is more amateur radio-oriented and addresses specific types of disasters. It gives straightforward guidance to

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ABOVE & BEYOND

VHF and Above Operation

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VHF to microwave: how to repair old beam antennas

What started out as a winter storm here in San Diego became a severe path of destruction to my amateur-related activities and antenna system. High winds associated with the storm snapped my antenna guy wires and toppled my antenna. That was the start of all the fuss to reconstruct what had been my modest antenna system for VHF.

Evaluating the initial damage was a little hard, as the mast collapse (because of the snapped guy wire) could have been prevented by some simple maintenance. I had apparently left the system in the air for too long, without providing some care and upgrading or replacement of worn parts.

Time had made me complacent in the maintenance of my antenna system. I had not given any attention to a problem which had been waiting to happen for some time. In hindsight, we all should take time to give our systems some attention, to prevent having to replace antennas in a similar fashion.

The main culprit in my failure was the guy wires I had used. Through time they had become weak from stretching and just plain age. A simple bend showed that the material had become less malleable than new material. That should have been a great clue but it went unnoticed. All the signs were observed when the antennas were lying on top of my roof—after the system's key guy wire had broken in the high wind.

At least the roof wasn't damaged when the system came

down that stormy evening. Nothing was done at the time, because we determined that no further damage could be done as long as the antennas were resting firmly on top of my roof that Friday evening. Saturday morning would tell the full tale.

I am describing this adventure in reconstruction to show the efforts I went through in an attempt to be cost-effective in making the repairs to the old existing antennas, twisted as they were. First, the careful job of removing the pile of antennas and cable and getting down to business.

On first evaluation, the two-meter and six-meter beam antennas took the worst of the storm's fury. Actually, the six-meter beam boom stopped the entire structure from becoming a tangled mass of scrap aluminum because the long boom end hit the roof and stopped the remaining antennas from becoming an intertwined mess of metal.

All was not well, as the toppling over was done in an 80-or-so mph high wind gust. It just twisted the mast after the guy snapped.

When the antennas came to rest on the end of the six-meter boom there must have been quite a shock to all of their basic structures. The gamma match of the six-meter beam shattered into a number of pieces, as did several elements of the six- and two-meter beam antennas. The RG-58 coax feeding one antenna parted, but the RG-8 (RG-213) coax held together and was at full strain holding the antenna, which was now bent over on the roof. In a way, it became part of the guy system when it went over.

Reconstruction of the system started with cutting the RG-8 coax and getting rid of old

cables that were in the way. They had been in the air for many years and showed signs of age on the outer jacket. Most bolts holding the antennas were rusted beyond use and, while some came off with a little oil and elbow grease, quite a few had to be twisted off until they broke. First lesson: Spend the extra cash and use stainless or galvanized hardware on your antenna. They will not rust like common cad-plated nuts and bolts. You will thank me when you have to work on your antenna in a less hostile environment.

Removal of the beam antennas was next, placing what was left of the two- and 3/4-meter yagi antennas on the roof, off of the mast.

The last main structure to be removed was the six-meter beam. The original mast structure was removed and replaced with a new mast for the day when the antennas would go back up. Now the quandary was whether to purchase new antennas or repair the old. Taking stock, I determined that I could purchase new aluminum tubing for the six- and two-meter beam elements and (galvanized) hardware to reattach them for about \$12.

The trick to pull off was how to reconstruct the gamma match of both the two- and six-meter feeds. The original gamma match systems were constructed with some sort of plastic and two sections of aluminum tubing forming a capacitor of sorts for the feed. See **Fig. 1**. How to repair this gamma match for these two bands remained to be solved. Would it be better to purchase new antennas or repair the existing? I examined the coax cable and the answer became very clear.

The coax cable was years past due for replacement. I confirmed this by testing the loss of a 50-foot section of RG-213 that had been up for a lot of years. I made a test on a new piece of RG-213 and found that the loss of the old piece was about 1.5 dB greater, compared foot for foot. The old cable had to be

replaced. Purchasing new cable would be expensive. I checked the loss tables and found that Belden 9913 might be the best from the cost-to-low-loss standpoint. Expensive, but well worth it.

At two meters, the loss of Belden 9913 was one half the loss of brand new RG-213 (RG-8 quality coax). The 9913 specs at 1.5 dB and the 213 at 3 dB at 150 MHz. At 450 MHz, the Belden 9913 really shines, as its loss is 2.6 dB versus 6 dB for RG-8 types. This is a no-brainer decision—purchase the Belden 9913. Because the 9913 is a hard line type of coax, mating it to a gamma match requires that it be brought out to terminal lugs to bolt to the gamma match sections.

The #12 solid conductor of the 9913 cable was a slight problem because it has a spiral-wrapped internal construction that provides low loss but also allows water to enter at the coax end if not properly sealed. I solved this problem by crimping on my bolt connectors and then sealing the end of the open coax inner section with a small dab of RTV. I then overwrapped the center section and braid with good quality electrical tape, making many twists, figure-eights, and crossed-over tape sections to add additional protection from the weather.

Someone suggested the I apply a layer of varnish or plastic spray to cover the tape. On the advice of Kerry N6IZW, we tried some plastic sprinkler pipe glue and overcoated the tape with the plastic cement. This seemed to be best, as it really made what seemed to me to be a bond with the electrical tape and the jacket of the 9913 coax cable. This looked to me to be the best method of sealing the Belden 9913. Additionally, when I mounted the cable on the gamma match I positioned the cable on top and ran the crimp-on connections on down, forming a sort of drip loop preventing moisture from easily entering the coax cable. See **Fig. 2**.

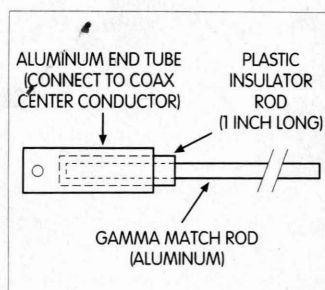


Fig. 1. Basic yagi gamma matching design typical of most commercial designs using plastic matching capacitor.

Now that my main expenditure was going for new Belden 9913 coax to replace my old lossy cable, the idea of a couple of new \$150 beam antennas for two and six meters would be scrapped and the old antennas repaired. With \$12 allocated for replacement element parts, the problem now remaining was how to repair the gamma matching scheme on the existing beam antennas.

Looking through various antenna handbooks and other literature, I found a simple rule stating that the gamma capacitor should be about 170 pF at 14 MHz and then scaled proportionally for other frequencies. Some simple math meant that at double the frequency, the capacitor value would be halved ($28 \text{ MHz} = 85 \text{ pF}$, $56 \text{ MHz} = 42 \text{ pF}$, making a two-meter capacitor somewhere in the 10 pF range).

I reconstructed my six-meter gamma match by soldering a 37 pF glass piston capacitor onto short sections of brass tubing that fit over the piston capacitor and would fit over the old aluminum gamma match adjustment arm. The end by the

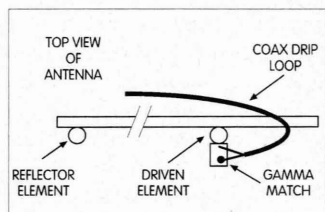


Fig. 2. Drip loop allowing rain to roll off the coax end termination by having the open end sealed and pointing down so it will not collect moisture.

connection to the insulator and center of the coax attachment was flattened and drilled to accept a #8 stainless bolt. The other end of the capacitor brass extension tube was slipped over the gamma arm and drilled to be secured with a 4-40 stainless bolt and nut. This attached the brass and aluminum rod firmly together, making a good electrical connection. With max capacitance at 37 pF, the brass sleeves were then overtaped and coated with plastic cement. See **Fig. 3**.

This worked amazingly well, but it soon became evident that the construction was not sturdy enough for the application. The glass piston capacitor and associated rebuilt gamma arm worked very well and was easy to adjust to resonance on the six-meter beam. However, this arrangement was not as good as I thought because I cracked three glass piston capacitors before I gave up on the idea.

In actuality everything worked well, but the glass piston capacitor I used was not sturdy enough. Looking through my junk box, I found several doorknob-type high voltage ceramic capacitors in the 25 and 100 pF range. These are the ones that have for the center connection a brass 6-32 screw mounting shaft.

I tried to insert the capacitors into the brass tubing I used with the earlier piston capacitors, and had to file off the edges. Then I soldered the capacitor onto the brass tubing sections and confirmed that while the value of capacitance was a little low, this worked quite well. The hammer test with a screwdriver proved the assembly to be quite rugged (see **Fig. 4**).

Because the assembly used a gamma capacitor of 25 pF, the gamma matching arm had to be set at maximum to bring the section into resonance at 50 MHz for a great SWR of 1:1 or quite near it. As the test was made on the ground, there were, I am sure, stray effects showing up—but it worked well.

Just as with the glass piston capacitor, 37 to 40 pF would be the ideal capacitance desired for

the ceramic high voltage capacitor as well. Junk boxes being what they are, I don't believe you or I will find a 40 pF capacitor. I had to settle on a 50 pF one and use it on the final gamma matching arm I constructed for the six-meter beam. The total cost of reconstruction for the gamma match was 50 cents and about five hours of elbow grease.

Oh, I forgot: There was that pint of gas to travel to K6DS's QTH to pick up the 50 pF ceramic capacitor that I used. I put the 50 pF unit on the antenna, confirmed a greater adjustment range over frequency, and then left it on for a final match when the antenna is up on the mast in the air.

For all these tests, I have attributed the ease with which I was able to make them to my MFJ-259 antenna analyzer SWR meter. Makes the evaluation and adjustment of antennas in the 2 to 150 MHz range quite easy.

Well, as you suspect, all the coax cables are cut to length and crimp connected in place, overwrapped, and sealed with connectors on the appropriate ends. What else remains to be done? In my case, it's mostly metalwork, making gusset plates to mount the boom to the two-inch mast section.

Here comes the hard part. I had to settle for less than I'd planned, as I could not find galvanized two-inch muffler U-bolts with back plates to fasten securely to the mast. I had to compromise with (untested) standard muffler clamps. These arrived as black steel and I am not sure how they will stand up to the weather.

I manufactured my mounting plates from four muffler clamps to secure the beam antennas to the two-inch main mast. It resulted in a very sturdy mounting arrangement, and was constructed out of scrap aluminum old panel sections that came out of the old junk box. I picked up the muffler clamps at a muffler repair shop. They came with the round-to-flat

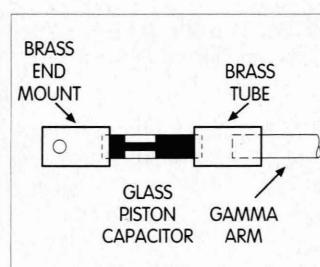


Fig. 3. The first attempt at constructing the gamma match section with a glass piston capacitor soldered to small sections of brass tubing for connections to the mounting insulator and gamma aluminum rod.

adapter clamp plate so necessary for securing the round tube of the boom and mast sections tightly.

The muffler-type clamps are probably available from some hardware source, but I was unable to locate them and settled for the muffler shop clamps. The cost was minimal for the clamps: one dollar for the two-inch ones, and less for smaller clamps. Quite reasonable. I lubed mine with a little grease and slipped an SMA coax protector cover over the bolt ends to protect them from rusting. Can't have everything, but you can prepare for the future.

I hope you have gotten some ideas on antenna reconstruction from my sad tale of the destruction that occurred on that Friday evening when things went "bump in the night." 73, Chuck WB6IGP. 73

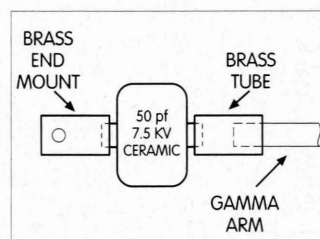


Fig. 4. Final gamma match capacitor construction using a 50 pF ceramic high voltage capacitor attached to brass tubing. Near end swaged for nut and bolt attachment and the other end inserted into the original aluminum gamma match arm. Attached with a 4-40 bolt through both brass and aluminum rod.

An RF Sensing Alarm

Save your power amplifier stages ... and your reputation!

Bertrand Zauhar VE2ZAZ
4176 Sylvio St.
Laval QC H7R 5V8
Canada

Have you ever walked into your shack and found out that one of your transceivers was on the air? Surprise! That has happened to me twice in the last couple of years. Once, it was the packet TNC that had crashed, leaving the two meter transceiver on the air for I don't know how long. That obviously blocked the entire packet BBS traffic. Whoops! The second time, my 70 cm satellite uplink transceiver and amplifier had been shooting 100 watts of RF at the stars for 30 minutes without my consent, thanks to one of the gizmos I had built to control the rig. Try to guess how hot the linear was! I decided that I'd had enough. Another occurrence and I would jeopardize my reputation forever.

I had to build a device that would alert me when it detected a continuous RF transmission lasting more than five minutes (a safe margin since I usually do not read the newspaper on the air!). The device would have to be broadband (HF/VHF/UHF), be sensitive enough to detect a five-watt transmission from inside the shack using a telescopic antenna, and produce a sound loud enough to alert me anywhere in the

house. It would also have to be self-contained, without any hookups to my radios. After a bit of reading and thinking, I came up with a solution that meets all the initial objectives. Here it is in detail.

Circuit description

The circuit shown in **Fig. 1** may look scary for some of you, but it isn't. It can be broken down into four stages. Let's look at them one at a time.

The first stage acts as an RF sensor circuit. It is made of U1C, one of the four operational amplifiers of an LM324 chip, and its associated input circuitry. U1C is used as a voltage comparator. Note that the two U1C inputs (plus and minus) have similar DC circuits connected to them. The plus input has R7, R8 and D3, and the minus input has R6, R10 and D5. In these two circuits, D3 and D4 are partly biased (about 200 mV of forward voltage) to better exploit the variation of voltage versus current that the diode produces. This translates into increased RF sensitivity.

In an idle condition (no RF detected), potentiometer R10 is set to make the voltage at the minus input of U1C slightly lower than that at its plus

input. This keeps the output of comparator U1C saturated to the "high" state (near supply V+). A strong RF signal present at the antenna terminal J1 reaches D3, a Schottky diode, through C3, a coupling capacitor. The diode rectifies this signal and generates a drop of voltage at D3 anode. This makes the voltage at the plus input of U1C fall below the one set at the minus input. As a result, the output of comparator U1C flips down to the "low" state (near ground). In this state, the circuit is in RF detection mode and timer U1B is activated. Note that as soon as the RF signal disappears, the comparator immediately returns to its high state, resetting the entire alarm circuit. Ferrite bead FB1 and resistor R8 are used to block the RF from reaching comparator U1C. Additional RF decoupling is provided by capacitors C2 and C5.

The second stage acts as a timer. It is made of U1B, D4, C4 and R9. U1B is again used as a voltage comparator. When no RF is detected, capacitor C4 is kept charged by the "high" state of U1C. When RF is detected by the first stage, capacitor C4 is left "floating" and starts to slowly discharge through R9. When C4 is sufficiently discharged,

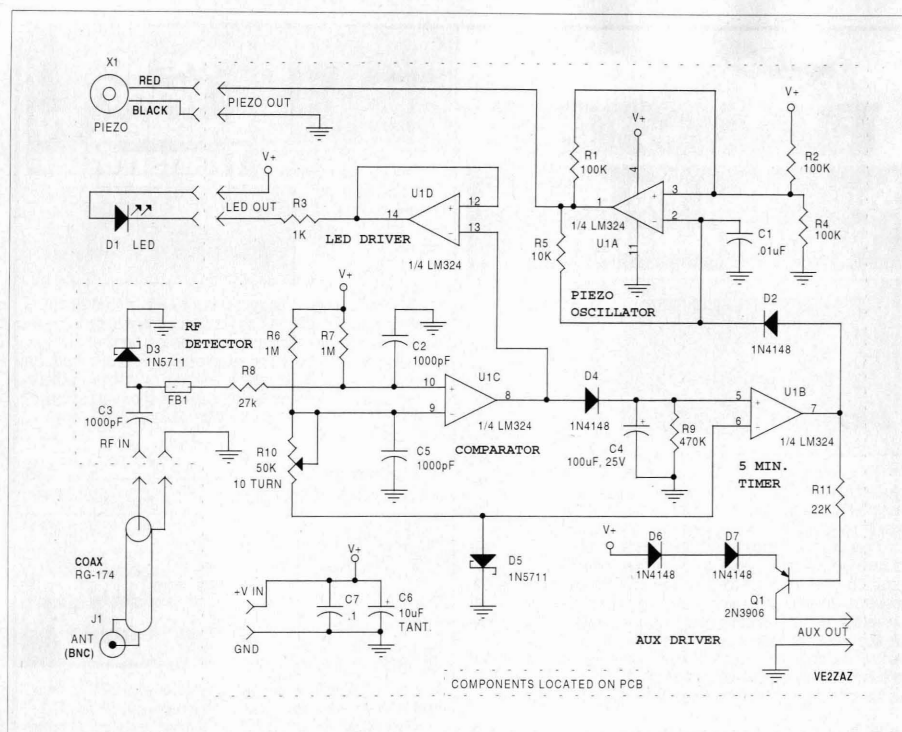


Fig. 1. Circuit diagram for the RF Sensing Alarm.

after approximately five minutes, the voltage at the plus input of U1B falls below the one at the minus input which is set by D5 (about 200 mVDC). As a result, the output of comparator U1C flips down to the "low" state (near ground). In this state, the alarm is tripped. The timer stage drives piezo oscillator U1A and PNP transistor Q1. The latter is added to drive an external device with a voltage close to V_+ when the alarm is tripped. The two diodes in series with the emitter provide a drop of voltage so that the base-emitter junction of Q1 is not biased when U1B output is in "high" state. The 2N3906 or equivalent transistor will safely supply a current of up to 50 mA.

The third stage is an astable multivibrator (square wave oscillator) and is made of U1A and surrounding compo-

nents. It drives the piezo transducer to produce a loud high-pitched sound. The oscillator circuit will operate only when the output of U1B presents a "low" state. Diode D2 serves as an isolating device between the two stages when U1B is in "low" state. The oscillator's frequency is set by capacitor C1 and resistor R5. The values chosen make the circuit oscillate at approximately 2600 Hz, a frequency that causes the piezo to generate the loudest sound.

The last stage is a simple buffer U1D that sinks current to turn on LED D1 whenever an RF signal is sensed by U1C. R3 limits the current to a safe level for the buffer, less than 20 mA.

The balanced input configuration of comparator U1C allows the supply voltage V_+ to vary over a wide range and the entire circuit will still work. I designed the circuit for a 13.8 volts supply, but I verified that the circuit works down to approximately 10 volts without recalibrating it, and down to approximately five volts if recalibrated. The high end of the range is more delicate to set since devices such as the LED, the piezo transducer and polarized capacitors may be overstressed if component value changes are not made. Consequently, I would not recommend going beyond 20 volts.

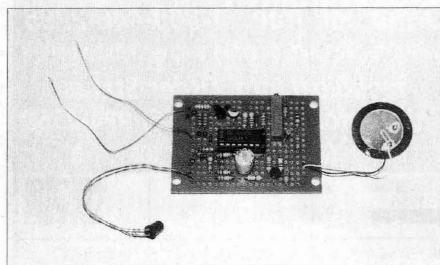


Photo A. The prototype, assembled on a universal printed circuit board.

Circuit assembly

This project is relatively inexpensive to assemble. If all the components are purchased, it should cost less than \$25 to build it. Obviously, your junk box's size will dictate the cost. I reused components taken from old PCBs and built the project for less than five dollars, including PCB and box.

As shown in **Photo B**, I elected to design a PCB for the project. It is a rather compact design. The intent here is to make the PCB fit in a common Hammond 1591A plastic box. Note that the circuit can be assembled using other techniques: universal PCB, veroboard, dead-bug, wire-wrap and even a combination of these techniques. The layout is not critical except for the RF portion of the circuit: Components D3, C3, FB1 and R8 should be mounted as close as possible to each other, using very short leads. This will guarantee proper operation up into the UHF spectrum. Also, decoupling capacitors C2 and C5 should be mounted as close as possible to pins 9 and 10 of U1C.

An IC socket for U1 is desirable since the LM324 is rather sensitive to pin shorting compared to other operational amplifiers I've used in the past. With a socket, replacing it is a snap.

Ferrite bead FB1 is inserted over a short piece of solid wire and the wire is soldered to the PCB pads. Potentiometer R10 is located on the PCB edge so that its adjustment screw can be accessed through a small hole drilled on one of the faces of the box. I've chosen not to put connectors for external connections to the LED, the piezo transducer, supply line and the auxiliary output. Solder small gauge (#26 or smaller) wires directly to the PCB pads and connect the other ends to the external devices.

For antenna connection, I recommend using a female panel-mount BNC connector. It is small, reliable and will accept most right-angled telescopic antennas when mounted horizontally. Other RF connectors can be used if desired. Since the input impedance of the circuit is quite high, maintaining a constant impedance through the connectors is not an issue. In general, though, it is good practice to

avoid using UHF connectors (PL-259/SO-239) when working above 200 MHz, since they are not of constant impedance type. For coaxial cable, a short piece of RG-174 type or equivalent is preferred to limit overstressing of the PCB pads. This type of coax is much easier to route in a small box anyway. Solder the bare ends directly to the pads.

Someone may decide that five minutes of margin before alerting is too short. C4 and R9 make up the RC circuit that determines the duration. The discharge rate can be varied by changing the value of C4 and/or R9. Fig. 2 shows a plot of resistor R9 as a function of time for different supply voltages V+. This graph gives a ballpark idea of the resistor value to pick. I computed this graph using a tantalum capacitor for C4. I verified that, in general, tantalum capacitors match closer to the theoretical exponential discharge curve than do electrolytic capacitors. If an electrolytic capacitor is used instead, the user can expect to get about 10 % longer duration than what is indicated on the graph. Electrolytic capacitors are far from being perfect devices (farther than tantalum) and their unwanted characteristics tend to lengthen the discharge duration. Also note that tolerances on the capacitor values will affect the duration as well.

Some comments

Piezo transducers have a resonance frequency that, when submitted to, produces the loudest sound. The piezo I used came out of my junk box. I found that a 10 k resistor for R5 produced the loudest, most unpleasant sound. Using

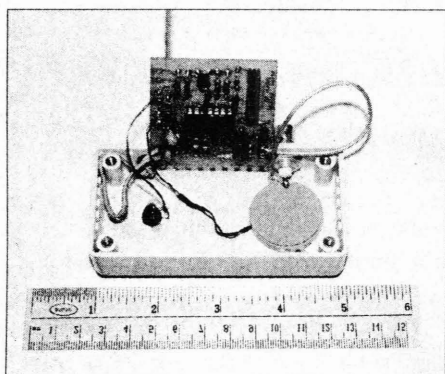


Photo B. The fully wired unit, before closing the cover.

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9117	17 meters		



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a different piezo element may require a different value of resistance. I suggest temporarily connecting a 50 kΩ potentiometer to find your piezo's "sweet spot." You can then measure the required resistance with an ohmmeter directly on the potentiometer and substitute it with the closest resistor value at the end.

You may have noticed that some piezo transducers have three wires and some have two. Only two wires are used here. The third wire is a feedback signal. If the piezo you are using has three wires, the black wire goes to ground, the red goes to the oscillator and the blue (or other color) is not connected.

Make sure you use an "external drive" type of piezo transducer. Some piezos come with built-in oscillator circuits and will only function if they are connected to the auxiliary output provided by Q1.

Finally, if means of alerting other than piezo are preferred, you can use the auxiliary output to drive an external device. In this case you don't have to install components R1, R2, R4, R5, C1 and D2. Put jumper leads at the place of C1 and R4 to ground the unused LM324 inputs.

Continued on page 14

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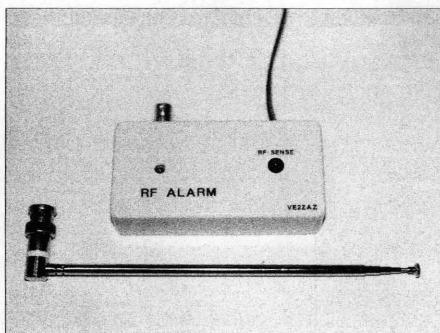


Photo C. The completed project and the telescopic antenna.

Calibrating and testing

When everything is assembled and visually inspected, it is time to apply DC supply to the circuit. When doing so, if the piezo sounds right away, adjust R10 until it stops. If you can monitor the supply current to the design using an ammeter, it should be less than 10 mA. Next do the touching test. There should not be any hot components.

Diodes D3 and D5 must be at the same temperature before calibration is performed. This is critical since temperature offsets can create millivolts of difference and unbalance the circuit. Do not touch the PCB for a few minutes before going any further. To calibrate the unit you will need a digital voltmeter of 10 M Ω or higher input impedance that can read down to the millivolt. First connect the voltmeter between U1C pin 10 and ground. Measure the voltage and make a note of it. Then move the voltmeter's red probe to U1C pin 9. Adjust potentiometer R10 to bring the voltage at pin 9 higher than the voltage previously measured at pin

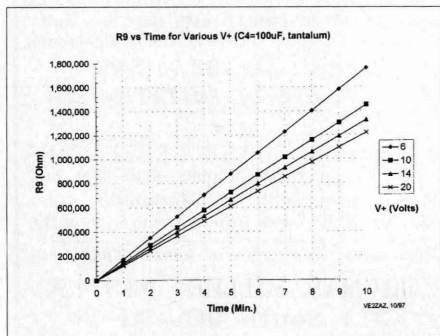


Fig. 2. R9 vs. time for various V+ (C4=100 μ F, tantalum).

10. The LED should turn on. Backing off on R10 to bring the voltage below the one at pin 10 should turn off the LED. This verifies that the comparator stage is working properly. Now, adjust potentiometer R10 so that the voltmeter reading on pin 9 is three to five millivolts lower than the value noted from pin 10. This completes calibration.

At this point you are ready to verify the circuit using RF. Attach an antenna to J1. Transmit a constant amplitude signal (AM, FM, modulated SSB) from any of your radios with at least five watts of output power (using another antenna, of course!) and verify that the LED comes on when RF is present. Moving the antenna around in the shack, for example close to a window, may give better results. To verify the alarm circuit, you must maintain RF presence for at least five minutes before you can hear the sound. Make sure the frequency chosen will not interfere with ongoing QSOs, and identify yourself often. When the alarm sounds, stop transmitting and verify that the alarm stops and that the LED turns off. Try detecting RF at other frequencies to verify sensitivity flatness.

Performance

Using a calibrated signal generator, I verified that the alarm's sensitivity is about seven millivolts RMS of RF and is quite flat from 1 MHz to 550 MHz. Sensitivity depends on how close the two voltages at the inputs of comparator U1C are. The circuit is stable as long as the entire design is at the same temperature. This is the case when all the electronics are enclosed in a box. I've used the alarm for a few months now and haven't had to recalibrate it. Since the RF sensing circuit is not tuned, selectivity is dictated by the type of antenna used. I found that about two feet of antenna is enough for proper operation at all frequencies, HF, VHF, UHF. I put the alarm box on top of my radio console and I use a telescopic antenna. Note also that my shack is located in a basement. A bit of experimenting here with antenna type, position and length may be required. I've also verified that I can hear the alarm sounding wherever I am located

Parts List

C1	.01 μ F ceramic
C2, C3, C5	1000 pF ceramic
C4	100 μ F, 25 V tantalum or electrolytic
C6	10 μ F, 25 V tantalum or electrolytic
C7	.1 μ F ceramic
D1	LED, any color
D2, D4, D6, D7	1N4148 or equivalent
D3, D5	1N5711 Schottky
FB1	Ferrite bead
J1	BNC, female panel mount
Q1	2N3906, PNP or equivalent
R1, R2, R4	100 k, 1/4 W 5%
R3	1 k, 1/4 W 5%
R5	10 k, 1/4 W 5%
R6, R7	1 M, 1/4 W 5%
R8	27 k, 1/4 W 5%
R9	470 k, 1/4 W 5%
R10	50 k, 10-turn pot (Bourns 3006 type)
R11	22 k, 1/4 W 5%
U1	LM324 quad op amp
X1	Piezo, MuRata PKM-11 or Radio Shack #273-73 or equivalent
Socket	14-pin DIP
Coax	RG-174 or equivalent
Box	Hammond 1591A suggested
Right-angle BNC telescopic antenna	

Table 1. Parts list.

in the house. This sound is equivalent to a smoke detector, so you shouldn't have any problems hearing it.

I have not been alerted by my new alarm yet, but it gives me a peace of mind that I never had before. Knowing that it's there makes all the difference! 73

Sensitivity Training

Some pros and cons of increasing receiver sensitivity.

Parker R. Cope W2GOM/7
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It seems we're always scratching for a little more sensitivity, but sometimes our scratching just makes a bad situation worse. A low noise amplifier (LNA) can help in special circumstances, and those are usually found above 15 meters. In the HF range below 20 MHz, the noise the signal is competing with is primarily man-made, but atmospheric noise produced by lightning discharges in thunderstorms can be a major problem depending on the time of day, season, weather, location, and frequency. The QRN on 160 meters and 80 meters in the summertime can be a bear. Man-made noise is an inverse function of frequency and depends on location, in that it is due chiefly to electric motors, neon signs, power lines, ignition systems, and the ubiquitous TV receivers' sweep generators. Above 30 MHz, man-made noise is about equal to thermal noise.

Thermal noise is generated in the thermal agitation of electrons in a resistance and is independent of frequency. The electron or hole flow in a transistor or the electrons flowing from cathode to plate in a vacuum tube are discrete units of charge occurring

randomly. Their sum makes up the DC current, and their RMS value makes up the AC value. The AC component exhibits all of the characteristics of thermal noise. The RMS variation of transistor current is difficult to predict, and only under special circumstances such as in temperature-limited diodes can it be defined statistically. Thermal noise power (watts) available from a resistor is expressed as kTB , where k is Boltzman's constant, 1.38×10^{-23} , T is the absolute temperature in degrees Kelvin (room temperature is about 300°K), and B is the noise bandwidth in hertz (essentially the 3 dB bandwidth). kTB has a value of -203 dBW or -173 dBm per hertz. In a 1 kHz bandwidth, the noise power is -143 dBm.

The maximum available power from a generator is obtained when the load resistance equals the generator's resistance. Therefore, the RMS value of the noise voltage from a resistor R is:

$$e_n = \sqrt{(4kTBR)}$$

The radiation resistance of an antenna produces thermal noise, which is the minimum signal available from the antenna. As far as the receiver is

concerned, noise from the antenna is just more signal. The noise the desired signal must compete with is the sum of the antenna's noise and the noise generated within the receiver. We can't do anything about the noise generated by the antenna, but the noise generated in the receiving system is a different story. We can increase the antenna's signal and noise to a level that overrides the receiver's internally generated noise. Generally, we are stuck with the receiver's internal noise, but some receivers are better than others.

The sensitivity of HF receivers is often given in terms of the RF signal at the input of the receiver needed to produce some signal-to-noise ratio S/N at the output. For example, an HF receiver may be specified as 0.5 μV for 10 dB S/N . VHF and UHF receivers usually specify the sensitivity in terms of "Noise Figure" or "Noise Factor," which is a comparison of the equivalent noise at the input of the actual receiver to the noise at the input of a perfect noise-free receiver when fed by a generator with a matching source resistance.

The noise factor of a network or "black box" can be defined as follows:

receiver, true enough, but these higher signal levels are more likely to overload the receiver and generate distortions that appear as false in-band signals. Of course, a step attenuator in front of the receiver can negate the preamp's effects as if the cable losses were increased. A resistive attenuator is recommended over gain control of IF or RF stages in the receiver because the attenuator is passive, linear, and located at the lowest signal point. Many HF receivers have a switched 20 dB attenuator at the input to ease the overload potential. The attenuator actually increases the dynamic range; the maximum signals the receiver can handle don't change, but the minimum signals are reduced to near the receiver's internal noise, so the ratio of max to min increases.

The step attenuator shown in **Fig. 1** has three 10 dB steps, for a total of 30 dB. The resistor values for a 50 Ω step attenuator are given in **Table 1**. For a 75 Ω attenuator, multiply the values given in **Table 1** by 1.5. Of course, additional attenuator sections or different attenuations can be used if desired. The maximum attenuation that can be obtained in one step is determined by the capacitive coupling across the switch contacts. With ordinary DPDT toggle switches, the maximum attenuation is less than 40 dB at 30 MHz. R_1 is the series resistance and R_2 is the input and output shunt resistance of any one attenuator section. An attenuator with a total attenuation of 30 dB maximum is probably adequate for most situations. If the attenuation is set so that noise can be comfortably heard with maximum receiver gain, you won't miss any signals and you'll have the maximum dynamic range.

Even good HF receivers often suffer from poor sensitivity on 10 and 15 meters; typical values are 0.25 μ V for 10 dB S/N (12 dB NF). A moderate preamp can overcome the sensitivity shortcomings of a poor 10-meter receiver. For example, a preamp with a noise figure of 4 dB and effective gain of 20 dB can improve the noise figure of a receiver with a 16 dB NF to about 4.6 dB. That's like multiplying the number of elements in your beam by nine or 10.

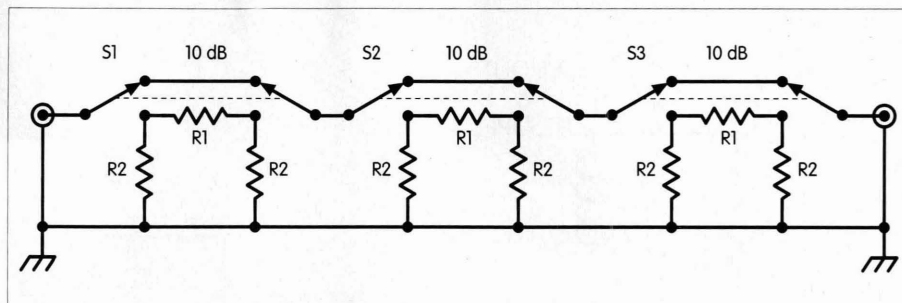


Fig. 1. A step attenuator is a coarse gain control.

A preamp can be built with discrete transistors or with hybrid amplifier modules such as the Motorola MWA110. While a preamp using a 2N5460 JFET can have a spot noise figure as low as 1 dB up to 400 MHz, obtaining a broadband match without introducing excessive losses is a challenge. There is much to recommend the MWA110 hybrid amplifier: wide bandwidth, respectable noise figure, and 50 Ω input/output resistance. The inherent stability of the hybrid modules makes possible the cascading of two or more units without oscillatory problems. An MWA110 is specified to have a bandwidth from DC to beyond 400 MHz, a gain of 14 dB, an NF of 4 dB, and 50 Ω input and output resistances. The 4 dB noise figure isn't great, but it's not bad considering the bandwidth. Two MWA110s in cascade have an overall NF of 4.1 dB and a gain of 28 dB. When a two-stage preamp made with MWA110s precedes a receiver with a 16 dB noise figure, the overall noise figure is about 4.3 dB. At 2 meters, where line losses are 7 dB and the receiver's noise figure is probably less than 10 dB, the overall noise figure improves to 4.2 dB or better.

The internal circuit of the MWA110 and the schematic of a single-stage amplifier are shown in **Fig. 2**. At VHF and above, the input/output impedance levels are most easily preserved on a circuit board by using 50 Ω microstrip transmission lines. **Fig. 4** is an example of a two-stage MWA110 amplifier which uses microstrip lines in conjunction with other sound RF construction techniques.

The construction of VHF circuits need not be difficult or magical if you keep in mind that parasitic capacitance

and inductance accompany every component, and that the resistance of conductors is increased by skin effect. These effects must be accounted for—ignore them at your peril. The parasitic capacitance and inductance are independent of frequency but, of course, their reactances are dependent on frequency. For example, a one-inch piece of #20 AWG copper wire has an inductance of about 20 nH; at 400 MHz, that's an impedance of about 50 Ω . The lead inductance of capacitors and their capacitance makes an effective series-resonant circuit: A 0.01 μ F disc ceramic with half-inch leads resonates at about 10 or 12 MHz; with sixteenth-inch leads, it resonates at about 30 MHz. A strip of copper has lower inductance than a round wire: A one-inch-long copper ribbon a quarter-inch by two thousandths of an inch (the cross section area of #22 AWG) has an inductance of about 12 nH. Skin effect

Continued on page 22

Atten. dB	Shunt Arm R2 (ohms) *	Series Arm R1 (ohms) *
3	294	17.8
10	95.3	71.5
20	60.4	249
* nearest 1% values		
dB	R2 (ohms) **	R1 (ohms) **
3	300	18
10	100	75
20	62	240
** nearest 5% values		

Table 1. Resistor values for a 50 Ω step attenuator.

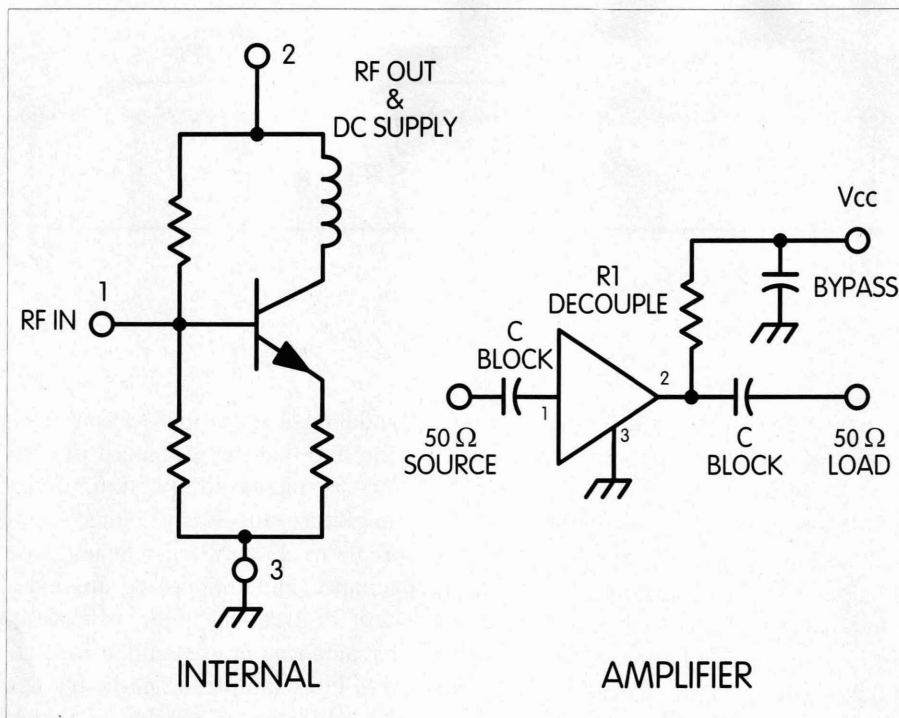


Fig. 2. The internal circuit of the MWA110 and its application as an amplifier.

Sensitivity Training continued from page 21

increases the resistance of a conductor when the conductor's inductance crowds the current to the outermost surface of the conductor so that the current-carrying cross-sectional area is decreased and the resistance increased. Skin effect is a function of frequency, permeability (μ) of the conductor, and the conductivity (σ) of the conductor. The critical depth δ where 63% of the current is concentrated is:

$$\delta = \sqrt{1 / 2\pi F \sigma \mu}$$

where

δ = critical depth in meters

F = frequency in Hertz

μ = permeability in Henrys per meter (for copper, μ is 12.56×10^{-7})

σ = conductivity in Siemens per meter (for copper, σ is 58×10^6).

These skin effects lead to DC ground not necessarily being RF ground. At RF, a ground plane should be used as opposed to a ground point. A ground plane is by definition a plane in which all points have the same potential: The inductance and resistance between two points on the ground plane is zero. At DC or low frequencies, the inductance or resistance of a wire can be essentially zero and a point can be the reference ground.

The power gain of the MWA110 is 14 dB when device current is 10 mA

and V_D is 2.9 V. When V_{cc} is 12 V, the decoupling resistance R_1 in Figs. 2 and 3 must drop 9.1 volts. Therefore, R_1 must be 910 Ω . The amplifier shown in Fig. 3 uses 50 Ω microstrip lines at the input and output of the MWA110. The DC components are on the ground plane side of the board, while the RF components are on the other side. The DC blocking capacitors C_1 isolate the amplifier from DC voltages on the input and output. C_1 and the 50 Ω input resistance act like a high-pass filter whose low cutoff frequency is $1/2\pi 50 C_1$. When the C_1 s are 220 pF, the low-frequency cutoff is 14.4 MHz; when the C_1 s are 1000 pF, the low frequency cutoff is below 3.5 MHz. C_1 is a leadless 220 pF chip capacitor similar to Kemet p/n C0805C221J5GAC. C0805C102K5RAC can be a 1000 pF chip capacitor C_2 that bypasses the higher RF current in R_1 to the ground plane. C_3 is a 0.01 μ F chip capacitor similar to Kemet C0805C103K5RAC that bypasses the lower frequency currents in R_1 to the ground plane.

Each C_1 causes the power gain to fall 3 dB per octave below the cutoff frequency, 6 dB per octave. When all three C_1 s are 220 pF, the response is down 9 dB at 14.4 MHz and down 18 dB at 7.2 MHz. Man-made noise rises about 20 or 30 dB in that octave. The increased noise and falling preamp gain will just about cancel out, and noise to the receiver will remain about the same. The resistors R_1 are 1/4 W carbon composition equivalent to MIL R-11 RC07GF911J. The R_1 s are mounted on the RF side of the circuit board between the microstrip near pin 2 to the V_{cc} patch. R_1 should be dressed down on the circuit board and soldered to the microstrip and the V_{cc} patch. Through-holes are not needed. L_1 and L_2 are Ferroxcube shielding beads 56-590-65/4A on #26 AWG wire. L_1 is a single turn and L_2 is two turns.

In passing, note that every pass of the wire through the bead's hole is one turn. The circuit board is one ounce (0.0014 inches thick) double-sided copper on 0.0625-inch Fiberglass™ epoxy (dielectric constant of 5), with the top side, which would normally be the component side, being the ground

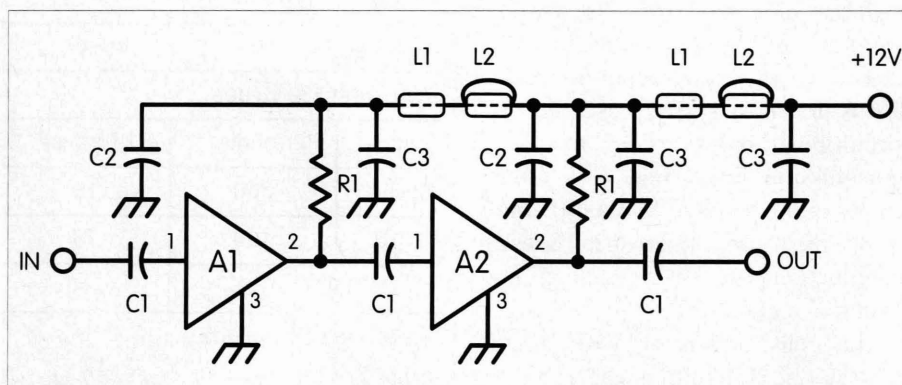


Fig. 3. Two MWA110s produce a broadband preamp.

plane. The ground plane is continuous except for reliefs around pins 1 and 2 of the MWA110s and around the through-holes for L_1 and L_2 . The bottom (track) side is the RF side. The top and bottom ground areas are interconnected with #20 AWG hairpin-shaped "U" bars soldered to both top and bottom ground areas. Of course, plated-through holes can be used instead of "U" bars. The number and location of the "U" bars is not particularly critical, but prudence indicates their use near pin 3 of the amplifiers, the bulkhead connectors, C_2 , and C_3 to ensure that the ground areas are all at the same potential. The inductance of a "U" bar is about 0.2 nH, and its reactance is less than an ohm even at 400 MHz.

The characteristic impedance Z_0 of the microstrip lines is a function of the circuit board's dielectric constant, thickness, and conductor width. A Z_0 of 50 Ω is obtained with one-tenth-inch conductor widths on the 0.0625-inch circuit board. The width of the striplines is critical, but not the length. Sharp bends in the lines can introduce discontinuities in the impedance and are better avoided. The MWA110 is mounted snugly down against the ground plane side of the board, and through-holes pass the leads through the ground plane to the microstrip. The case of the MWA110s is connected to pin 3 and may make contact with the ground plane, but it need not be soldered.

The schematic of a two-stage preamp using MWA110s is shown in Fig. 3. It is very much like the amplifier shown in Fig. 2, except for the decoupling in the V_{cc} line. Fig. 4 shows the stripline and the physical layout of the circuit board; the scale is approximately 2:1.

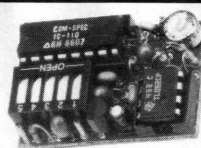
Continued on page 24

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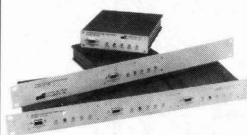
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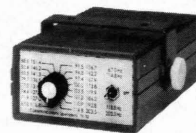
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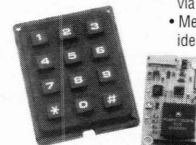
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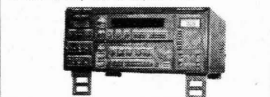
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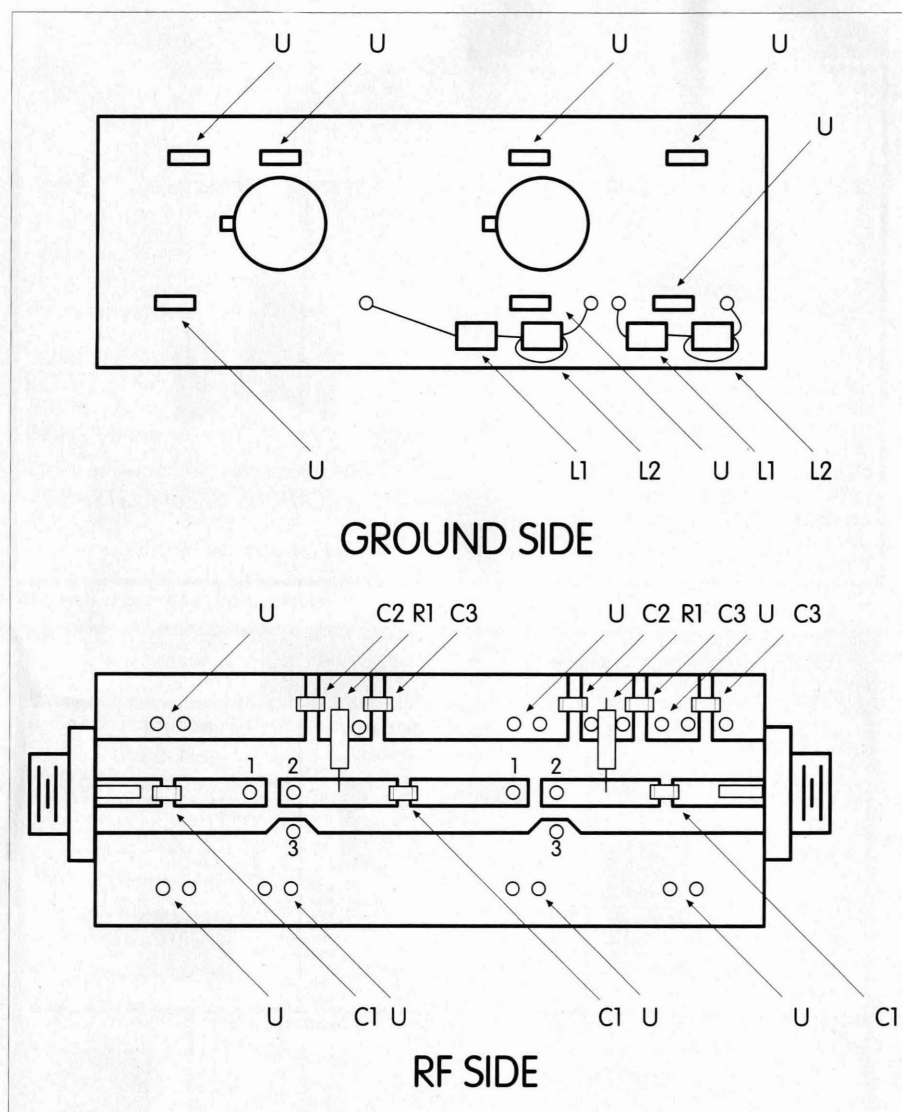


Fig. 4. A two-stage amplifier can be built in stripline.

Sensitivity Training

continued from page 23

When the amplifier modules are spaced an inch or so more apart, no interstage shielding should be necessary. The preamp

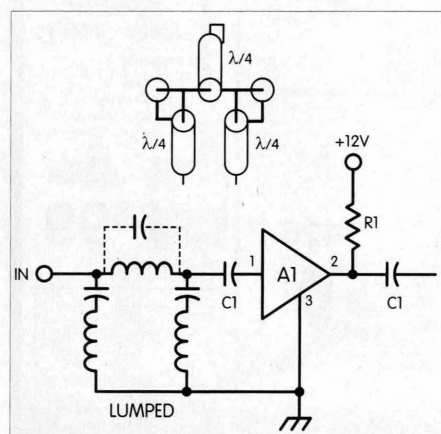


Fig. 5. A notch filter can reduce interfering signals.

board should be mounted in a shielded box and the length of the circuit board made long enough to fit between the center pins of the bulkhead connectors. The flanges of the bulkhead connectors should be soldered to the ground areas of the circuit board.

The shield box is necessary to protect against the weather and extraneous signals that might be picked up. Note that the enclosure can look like a waveguide or cavity and support waveguide coupling between the input and output as well as provide a means of undesired feedback. The coupling is not likely to be tight enough to be a problem if the connector center pins are soldered directly to the stripline and the decoupling resistors R_1 are dressed against the circuit board. +12 V should be introduced to the circuit board at the output end of the board. The board

is light enough to be supported by the center pins of the connectors and no extra mounting support should be needed for fixed installations. When used in a high-vibration environment, some extra support may be needed.

The advantages of a preamp at VHF and UHF are apparent when there are high transmission line losses or when the receiver's noise figure is high. The broadband preamp described can be used for all bands up to 440 MHz without band-switching. If there is a strong FM or TV signal to contend with, a notch filter will be required in front of the first stage to knock the big interfering signals down to tolerable levels. The filter shown in Fig. 5 can be made in coax between the antenna and the preamp or with lumped constants on the circuit board. The input impedance of a quarter-wave transmission line is an open circuit when the output end is shorted, and is a short-circuit when the output is open.

A typical lumped-constant filter is shown in Fig. 5. The tuned circuits resonate at the frequency of the interfering signal. For frequencies below 150 MHz, the quarter-wave lines are longer than 19 inches. For lower interfering frequencies, lumped constants are more appropriate. The inductor for the parallel-tuned circuit can be wound on a 1/4 W 10 kΩ resistor used as a coil form. Ideally, the inductor of the parallel-tuned circuit is self-resonant at the frequency to be rejected, and the shunt input and output series-tuned circuits are series-resonant capacitors. If wound coils are used, their axes should all be mutually perpendicular.

Most HF receivers don't shine at 10 and 15 meters, and a preamp can help there. But below 20 meters, the extra gain is no help and can actually be a hindrance. In the preamp shown, the gain below 20 meters is reduced by the DC blocking capacitors C_1 . When the C_1 s are 220 pF, the preamp gain begins to roll off at about 14 MHz where the extra gain is no advantage. A receiver's sensitivity can be improved with the judicious application of a low-noise preamp. The key word is judicious. Mae West wasn't talking about preamps when she said, "Too much of a good thing is just enough."

We Must Be Dreaming

A home-brew receiver project!

Thomas K. Duncan
1107 Alta Vista Ave.
Huntsville AL 35801

With last summer's hamfest season promising all sorts of new occupants for the parts bin, I felt it was time to make room by cleaning out some of the previous several years' good finds. Homebrewing several HF receivers with Gilbert cell front ends taught me the

perils of 40 meters—QRPers and half-megawatt broadcasters sharing the same band certainly tests dynamic range. As an SWL, I wanted to receive both. A receiver that would cover, say, 6.5 to 7.5 MHz seemed like both a challenging project and a way to clear space for this year's hamfest-find-of-the-century.

The circuit

Fig. 1 is the block diagram. Overall gain requirements were based on a $0.5 \mu\text{V}$ signal at the antenna delivering 0.25 W into an 8Ω speaker, or 134 dB. Originally, the intent was to use OH2GF's synchronous detector for AM (*A Synchronous Detector for AM Transmissions*, Jukka Vermasvuori OH2GF, *QST*, July 1993). This requires 15 mV input into 50Ω , so 90 dB gain is required before the detector, and 44 dB in the detector and audio sections. Ultimately, difficulty obtaining the NE604 or a suitable single-chip substitute in a standard DIP package led me to use a diode detector for AM and adapt OH2GF's product detector for CW and SSB.

The circuit is dual-conversion to improve image rejection. Double balanced diode mixers are used to generate the 10.637 MHz first IF and 455 kHz second IF. Ceramic filters are used between the first and second stages of each IF train. The MMIC first stages make up for mixer and filter losses, and provide 50Ω terminations for the mixers. The AGC-controlled second stages provide the bulk of the gain.

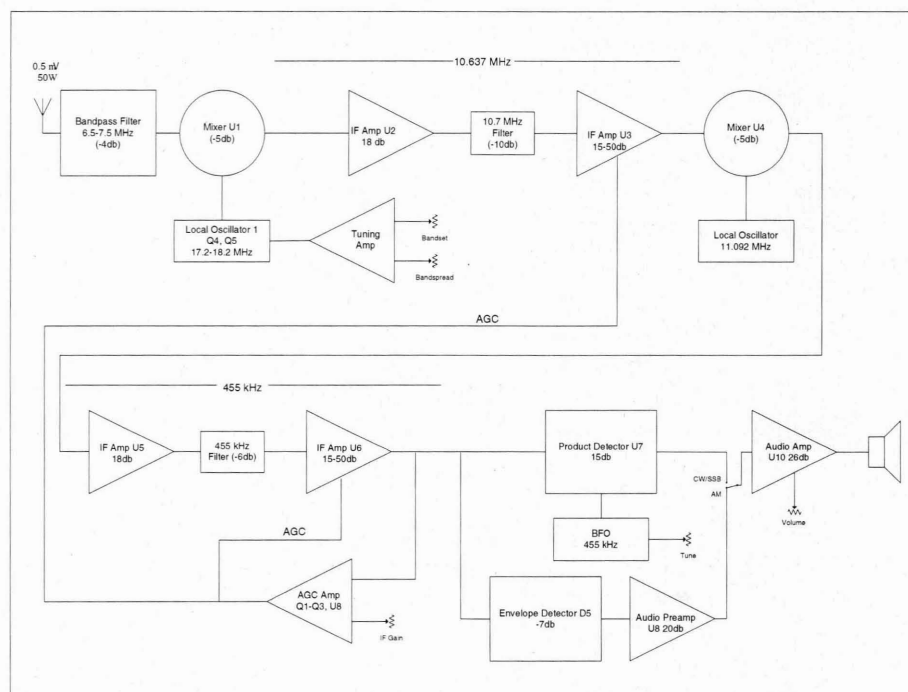


Fig. 1. Block diagram.

The first local oscillator and BFO are both varactor-tuned. The BFO tuning voltage range is determined by two 10 k trimpots, with which my junk box still overflows. First local oscillator tuning voltage is developed through what may seem to be a needlessly complex circuit. The bandspread control is a 10-turn pot with nice smooth action, begging to be pressed into service, but at 100 Ω and 1/8 W, it cannot by itself supply a wide enough voltage difference to tune the varactors over the desired 1 MHz range. An op amp is therefore used to multiply the difference between the bandset and bandspread wiper voltages to produce between 2 V and 5 V at its output, corresponding to between 17 pF and 11 pF of tank capacitance, for a frequency range of 17.2 MHz to 18.2 MHz. The buffer amplifier adapted from a circuit in Hayward and DeMaw (*Solid State Design for the Radio Amateur*, Wes Hayward W7ZOI and Doug DeMaw

W1FB; first local oscillator adapted from LO and BFO circuits for the 160 meter receiver in Chapter 6) isolates the oscillator and matches it to the 50 Ω first mixer local oscillator port. L9 adjusts the match to provide 0.5 V_{rms} (+7 dBm) injection.

The second local oscillator is a Pierce circuit using a readily available inexpensive crystal. At 11.092 MHz - 455 kHz = 10.637 MHz, the first IF is slightly off the usual 10.7 MHz, but the bandwidth of the first ceramic filter easily accommodates this. Similarly, the 10.7 MHz transformer has sufficient tuning range to adjust for a peak oscillator output of around 0.5 V_{rms} into the second mixer local oscillator port.

AGC is developed by amplifying second IF output through Q1 and Q2, rectifying through D6 and D7 and filtering the signal, and amplifying the resulting control signal through Q3 (see 1993 ARRL Handbook for Radio

Amateurs; AGC derived from Chapter 12, Fig. 43). The unity gain op amp functions as an inverter and level shifter, where the IF gain control sets the quiescent AGC voltage at pin 1 of U8 to between 4.5 and 6 V. As the second IF output increases, AGC voltage increases, reducing the gain of both MC1350s. AGC may be turned on and off by a push/pull switch mounted on the IF gain control, allowing IF gain to be adjusted manually even when AGC is used.

The other half of U8 is an audio preamplifier used to boost the envelope detector output to something near the output of the product detector. U7 has a conversion gain of around 15 dB, while the diode detector has a loss of around 7 dB. Most full-carrier AM signals in this receiver's range are broadcasters, so the 20 dB gain of the preamplifier is quite sufficient. Audio output is provided by U10, which

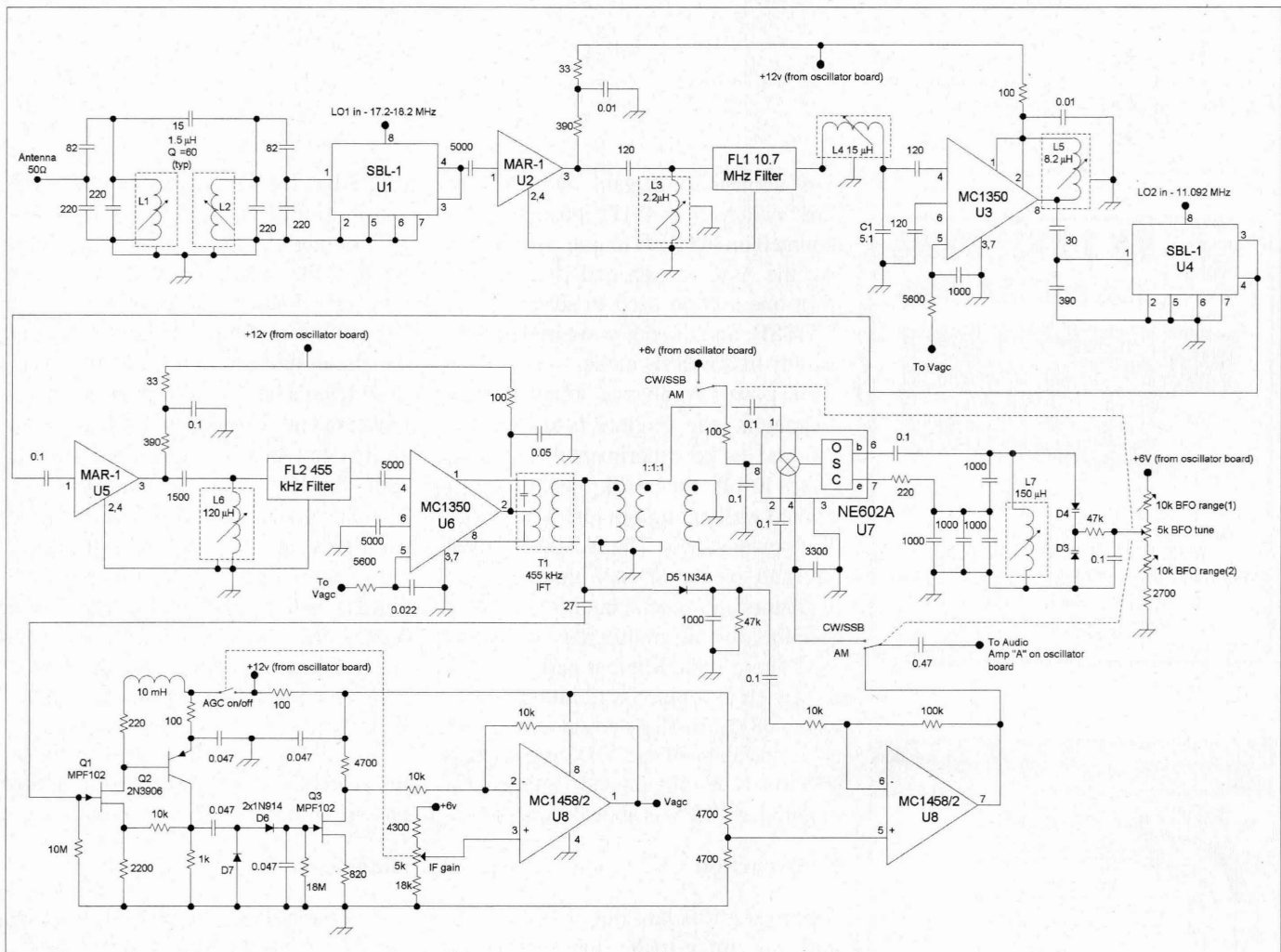


Fig. 2. Front end board schematic.

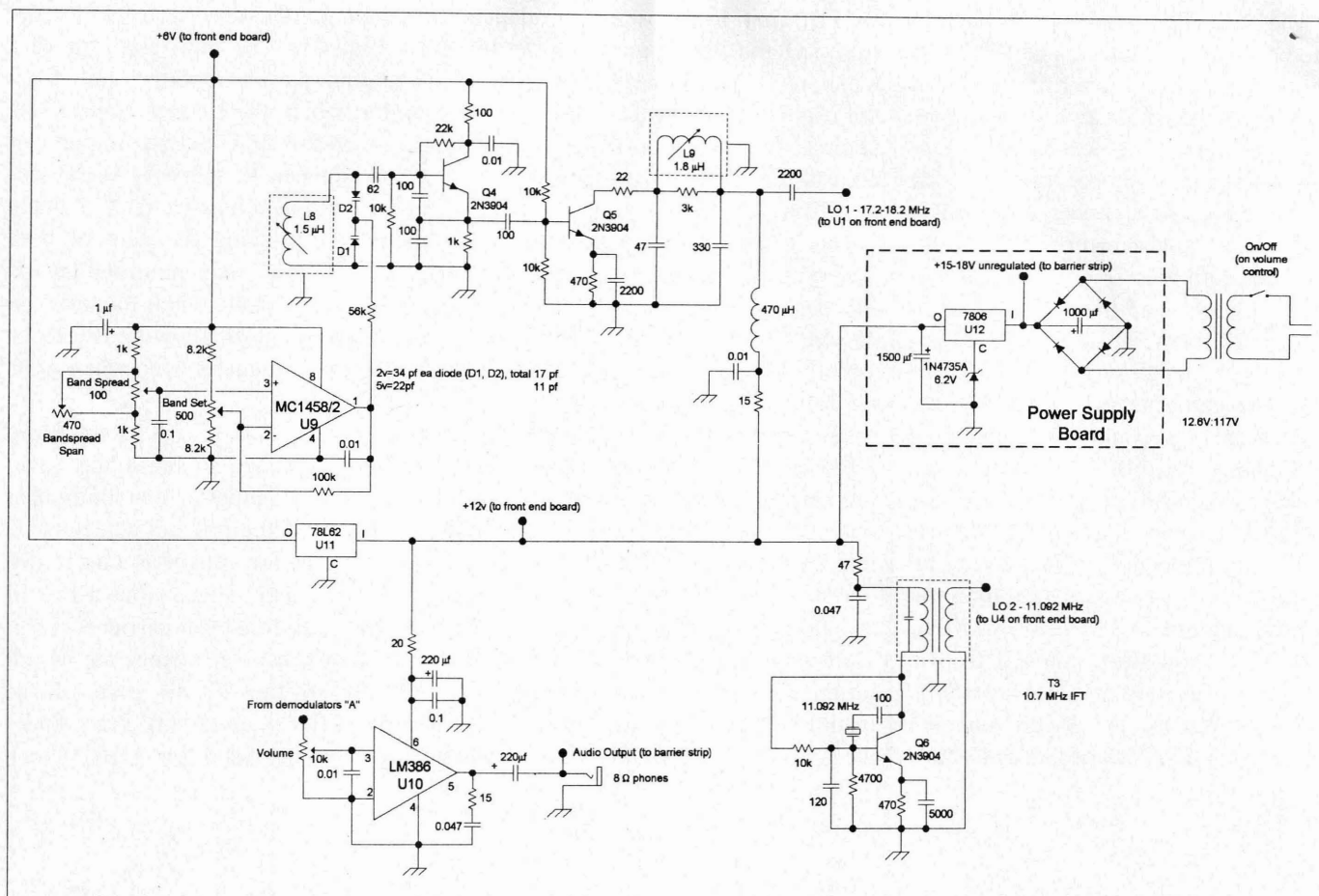


Fig. 3. Oscillator board schematic.

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The power supply was added as an afterthought—the original board set was intended to be experimental and never make it off the work bench, but it worked well enough to justify a permanent power supply. The fact that the junk box had plenty of 6 V three-terminal regulators and zeners, but no 12 V units, accounts for the multiplicity of regulators. Those with different parts on hand may wish to replace U12 and the zener with a 7812. Auxiliary power connections allow operation off +12 V DC regulated, or +15 to +18 V unregulated. Total current required at +12 V is about 90 mA.

Construction

The circuit is laid out onto two five-inch by three-inch single-sided PC boards. The oscillator board shown in

Fig. 5 has the first local oscillator and tuning amplifier, the second local oscillator, and the audio power amplifier. Each local oscillator section is partially shielded. The front end board shown in Fig. 4 holds the remainder of the circuit and has no overall shielding. Both boards have a good deal of unused space, but not so much that they could be easily combined onto one five-by-three-inch board. Most of the coils are shielded, so the circuit is well behaved even with the oscillator shields removed.

A three-inch by five-and-three-eighths-inch by six-inch box houses the boards and controls, with enough room left over for the power supply perfboard. Alas, the power transformer would not fit inside, so it's mounted on the outside of the rear panel. A barrier strip on the rear brings out the antenna, audio, ground, and auxiliary power connections.

Alignment

It is possible to roughly align the radio by tuning the two IF transformers "by ear" until some usable signal is

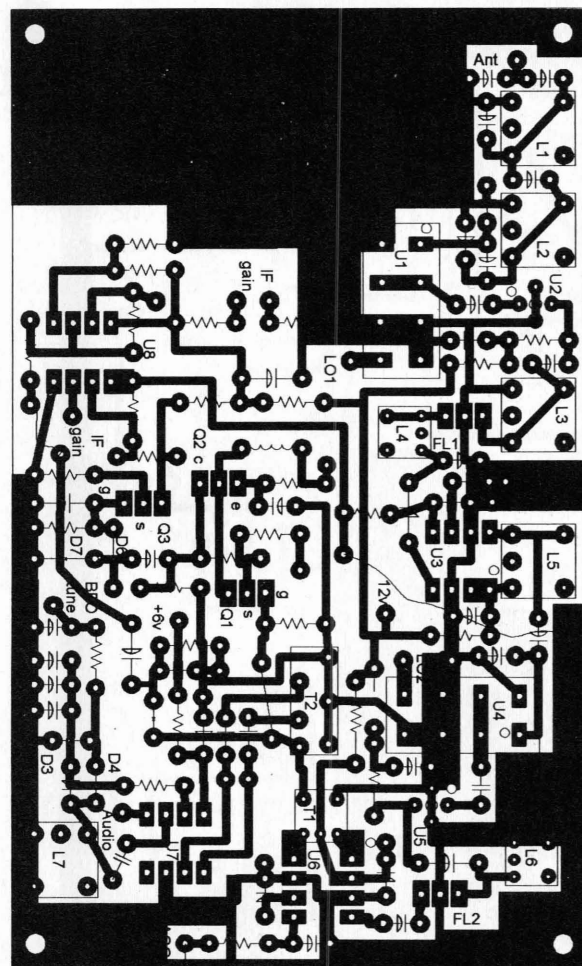
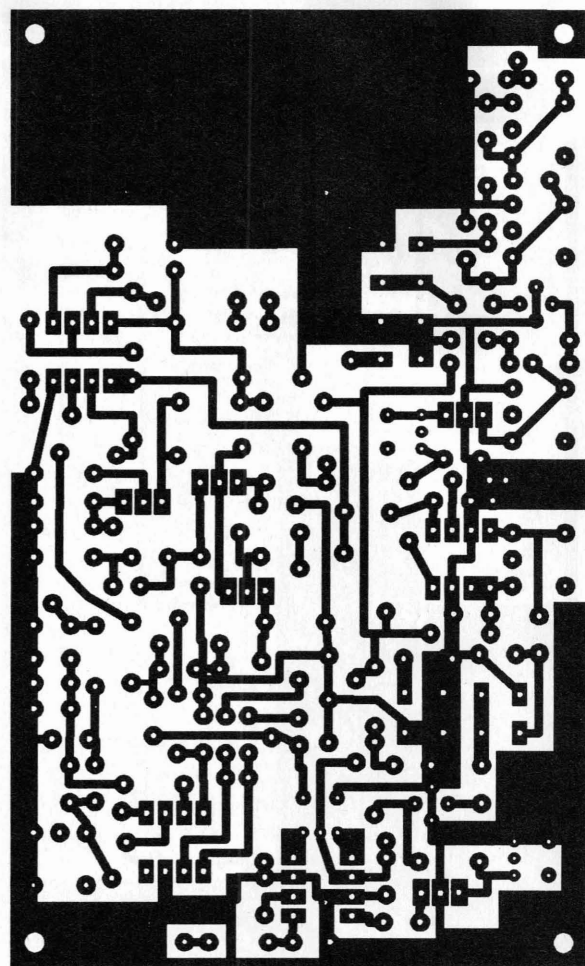


Fig. 4. (a) Front end board. (b) Parts layout from foil side.

found, and then adjusting L6, L5, L4, L3, L2, and L1 for peak signal. A more exact method follows, requiring a signal generator covering 7 MHz, voltmeter with RF probe or some other way of measuring RF voltages up to 20 MHz, and frequency counter.

With bandset and bandsread controls at center, the first local oscillator center frequency should be set to the center of the tuning range, 17.7 MHz, by adjusting L8's slug. L9 should then be adjusted for maximum injection into U1 pin 8, which should be around $0.5 V_{rms}$. Adjust T3 for maximum second local oscillator injection into U4 pin 8, again around $0.5 V_{rms}$. BFO frequency is centered by setting the two BFO range trimpots to their maximum values, centering the BFO tune control, and adjusting L7 so the signal at U7 pin 7 is at 455 kHz. The voltage at the wiper of the BFO tune control now corresponds to 455 kHz. The trimpots

are then adjusted to give the desired range (say, 1500 Hz) on either side of 455 kHz.

Attach the signal generator to the antenna terminals and insert a 1 mV 7 MHz signal, reducing signal level as you proceed with alignment and overall gain increases. You should be able to find this signal with the bandsread and bandset controls. Attach both the counter and voltmeter to the secondary of T1. Rock the bandsread back and forth and adjust T1 for peak voltage near 455 kHz. Proceed backward toward the antenna, adjusting L6, L5, L4, and L3 for peak signal. With the exception of L5, these are elements of low-Q tuned circuits. Adjustment will have very little effect and is not critical.

L1 and L2 establish the shape of the input filter. You may wish to peak both coils to favor a particular frequency, or adjust them so one peaks

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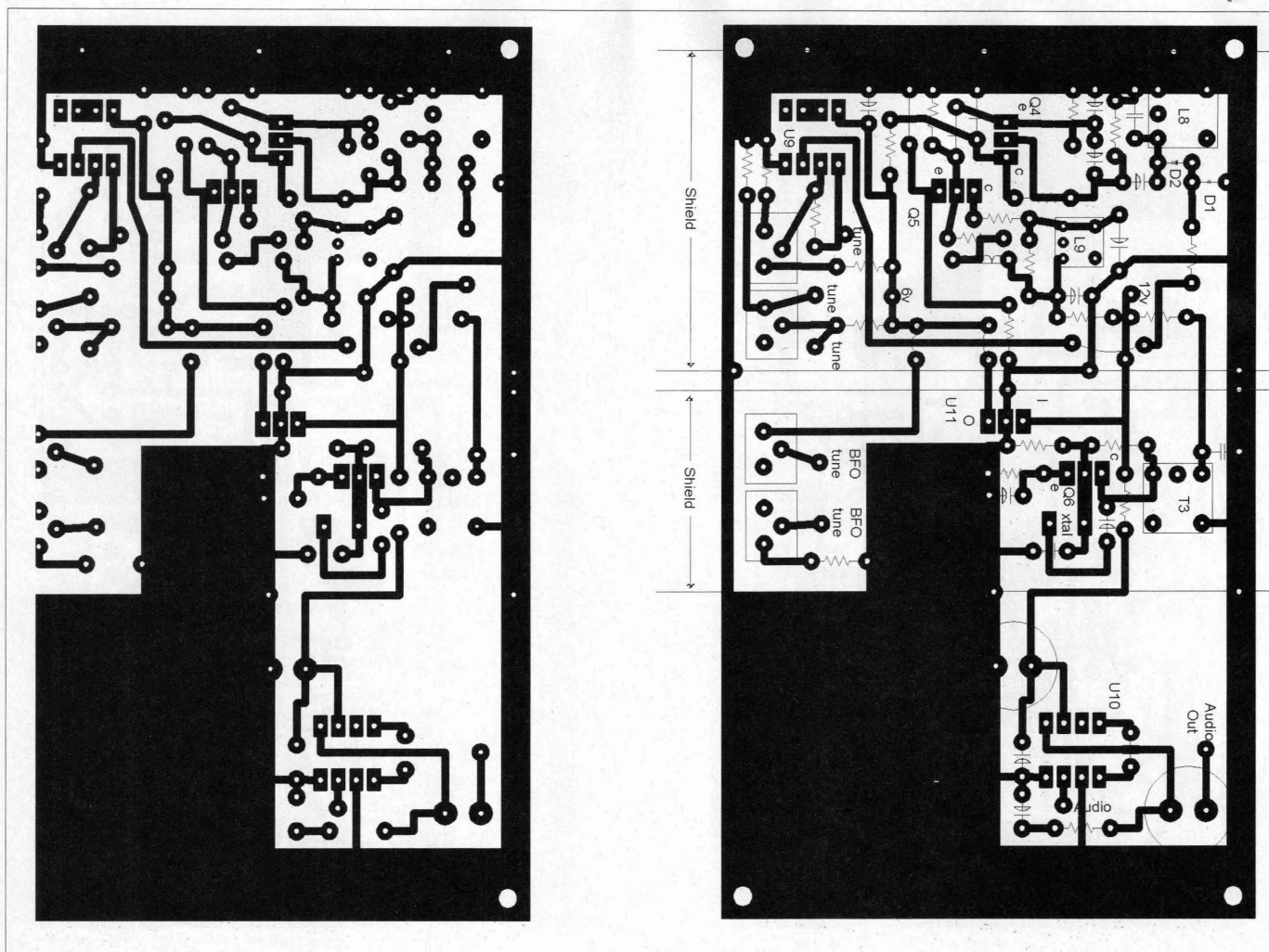


Fig. 5. (a) Oscillator board. (b) Parts layout from foil side.

below the center of the tuning range, and one above center, giving a flatter response.

Antenna

My unit was first tested on Field

Day, so there were plenty of signals to practice with. Even with plenty of signals (and don't forget, lots of atmospheric noise), a short length of wire on the antenna will not suffice. At 16 feet of elevation, I have a dipole cut

for 6.5 MHz, and a 75-foot random wire. Both these antennas, plus a cold water pipe ground, worked well.

Operation

The bandset control can tune AM signals by itself if you have patience, but then that's what the bandspread control is for. I have my bandspread span trimpot adjusted to give about 80 kHz/revolution, which is fine for AM. Decrease bandspread span resistance, giving a smaller tuning range, if you concentrate on CW and SSB. When the BFO is on, set the BFO tune pot to center and locate the signal, then adjust for the desired sideband and audio quality.

I generally set audio gain to about one-third, turn on AGC, and use the IF gain control (which functions even with AGC on) to adjust overall gain. The AGC responds to severe atmospheric noise in an annoying way, so in

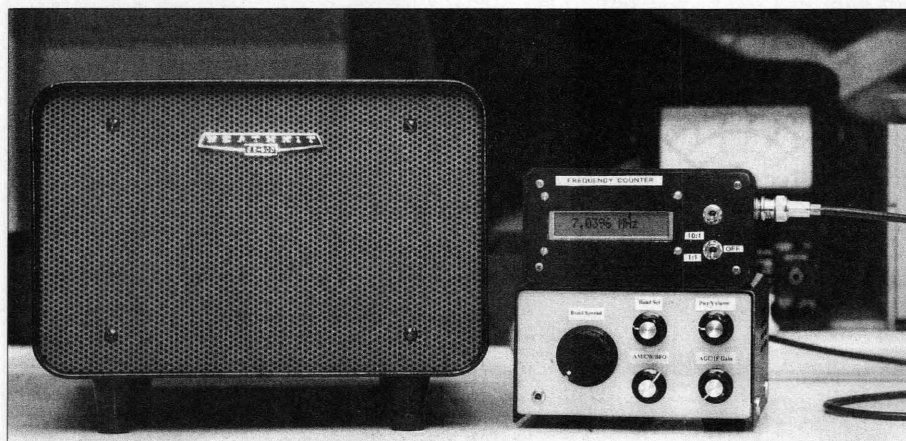


Photo A. Receiver shown with frequency counter and speaker. Note how the speaker, which normally accompanies my Heathkit SB-310 receiver, dwarfs both the radio and frequency counter.

that case I turn it off. There is plenty of overall gain, and enough audio to drive a speaker quite nicely.

Modifications

If you adhere to my "use the junk box" rule, you may wish to change the circuit to match your own hoard of parts. The first local oscillator, for example, could use a nice vernier-driven variable capacitor in lieu of the

Parts List

D1, D2	MV209 29 pF varactor
D3, D4	MVAM108 500 pF varactor
D5	1N34A germanium diode
D6, D7	1N914 silicon diode
FL1	10.7 MHz ceramic filter, 230 kHz bandwidth
FL2	455 kHz ceramic filter
L1, L2, L8	1.5 μ H slug tuned
L3	2.2 μ H slug tuned
L4	15 μ H slug tuned
L5	8.2 μ H slug tuned
L6	120 μ H slug tuned
L7	150 μ H slug tuned
L9	1.8 μ H slug tuned
T1	455 kHz IF transformer
T2	13 trifilar twisted turns on FT-37-77 core
T3	10.7 MHz IF transformer
U2, U5	MAR-1 MMIC wideband amplifier
U1, U4	SBL-1 double balanced mixer
U3, U6	MC1350 RF/IF amplifier
U7	NE602A mixer/oscillator
U8, U9	LM1458 dual op amp
U10	LM386 audio amplifier
U11	78L62 low power +6 V regulator
U12	7806 +6 V regulator

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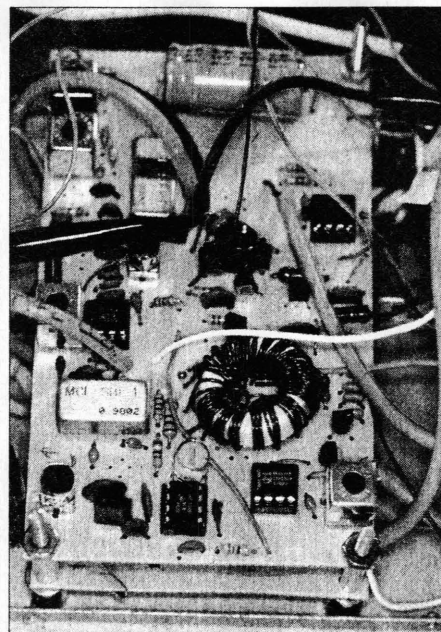


Photo B. Inside view. Large toroid is T2.

varactors, thereby entirely eliminating U9 and the bandset control. The same is true for the BFO. Both these modifications would improve frequency stability, which is a little wild for the

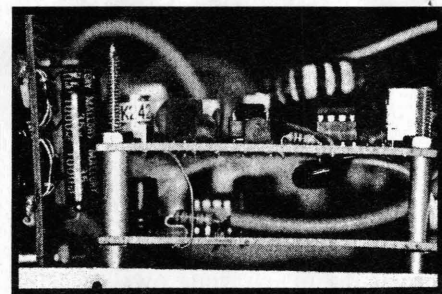


Photo C. Inside view showing stacked PC boards and power supply perboard.

first couple of minutes, but very solid thereafter.

There is enough space on the front end board to accommodate all of OH2GF's demodulator. The biggest component, T2, is already there. The SWL in particular should read OH2GF's article cited above before going ahead with the whole circuit or an AM-only version. Don't plagiarize my excuse about not being able to find an NE604!

L3, L4/C1, and L6 are impedance matching components. The circuit has sufficient gain that these components can probably be eliminated entirely. The π -section filter using L9 should not be removed, though: U1 and U4 like to see 50 Ω at each port. Speaking of filters, replacing the front end filter with one centered on 28.4 MHz would allow the low end of 10 meters to be tuned, with the first local oscillator used as is, but tuning on the low side. A single-stage RF amplifier might be needed to get the noise figure to a reasonable level.

There is only one second IF bandwidth: 6 kHz, determined by FL2. This is fine for AM, but a little wide for SSB and very wide for CW. An audio bandpass filter could be built from another op amp, and U9 has a spare section that could be used for this purpose.

73

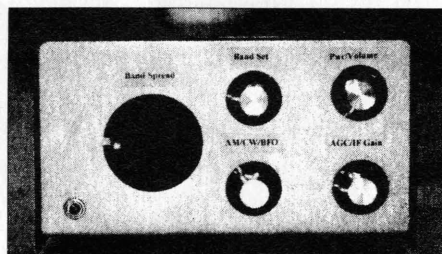


Photo D. Front view.

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10 GHz fun, 1999 update: the Ramsey FR-10 receiver

Quite a few years ago, I wrote an article covering a 30 MHz receiver using a single chip receiver from Signetics, the TDA7000. The purpose of this receiver and associated modulator was to provide a simple and effective method to get started on 10 GHz with wideband FM communications using surplus Gunn diode oscillators. The Gunn diode oscillators were obtained from garage door openers and from burglar alarm circuitry.

The conversion of the burglar alarms and the door openers was mainly to toss away everything except the metal cavity containing the microwave oscillator and detector diode. While these units are not as plentiful today, they can still be located in surplus, either at swapmeets or by scrounging in electronics junk bins.

Another source for Gunn diode units is SHF Microwave Parts Co., 7102 W. 500 S., La Porte IN 46350 [FAX (219) 785-4552 or E-mail through their Web page at (www.shfmicro.com) for details]. Together, Alan and his wife Pierrett have put together a great source for Gunn diode transceivers of various models for both 10 and 24 GHz frequency ranges. Take a look on their Internet page for current material. The 24 GHz Gunn oscillator is available for \$58 with varactor control (± 170 MHz) at 24.150 MHz (also includes a detector). Quite a bargain for a new unit!

These devices are listed on the SHF Microwave Web page. The unit is pictured, and specifications tell you in detail what you are looking at. Other items related to microwave operation are also listed, providing a complete list of all they offer. Be sure to check out SHF

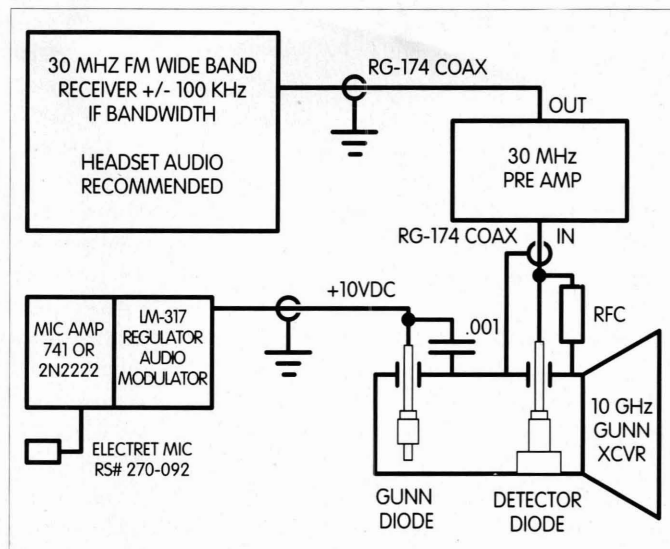


Fig. 1. Block diagram of 10 GHz system showing 30 MHz IF receiver with preamplifier, diode detector in cavity for receive, and Gunn diode in cavity for transmit. Modulator for Gunn diode is simple power supply voltage regulator, audio amp, and electret mike.

Microwave Parts Co. You'll be glad you did.

Available items include new Gunn-type oscillator detectors, and once in a while used items pop up. These units vary in complexity from simple Gunn diode oscillator detector diode units to those that are fitted with a third device, a varactor diode. The varactor diode allows for an easy frequency adjustment that is controlled by a DC voltage applied to it. While Gunn cavities that employ varactor diodes are more costly, they are much sought after for ease of operation

related to varying frequency of operation.

Gunn diode oscillator systems are inexpensive to start off with, using wideband FM rather than a more sophisticated SSB system. Well, the main difference among them is price and the test equipment needed to modify and test an SSB-type system. The cost for a working wideband FM system is about \$80, assuming that you purchase the components new. This differs from SSB-type systems, whose cost could run as high as \$250 for a kit of parts that you still have to modify from commercial frequencies to 10 GHz. There are fully assembled transceivers available from European manufacturers in a ready-to-use assembled unit for \$600 to \$1000, depending on RF power output.

While I operate on both wideband FM and SSB, I still recommend wideband FM for a beginning taste of microwave operation. As a matter of fact, I am in the process of constructing a 24 GHz wideband FM transceiver for use in the ARRL 10 GHz-and-up contest and to update with the Ramsey 30 MHz FM receiver system.

As I stated earlier, I began using a Signetics TDA7000 single-chip FM receiver tuned

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MAR 28

MADISON, OH The Lake County ARA will hold its 21st annual Hamfest on Mar. 28th, 8 a.m.-2 p.m., at Madison High School on Burns Rd. in Madison. The hamfest will feature new and used amateur radio, computer, and assorted electronic equipment, amateur-radio-related forums, an equipment test bench, and VE exams for those interested in earning an amateur radio license. Admission tickets are \$5 at the

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MACON, GA The Macon ARC will operate W4BKM 1500-2300 UTC on Sat. Mar. 27th, at the 17th annual Cherry Blossom Festival in Macon GA. Phone: 7.235, 14.240 and 21.335; CW 7.135, 14.035 and 21.135. For a certificate, send your QSL and a 9 x 12 SASE to Macon ARC, P.O. Box 4862, Macon GA 31208.

to 30 MHz for earlier transceivers. The modulator is nothing more than a voltage regulator that has its adjust terminal modulated with a low-level microphone-driven amplifier. The audio amp can be as simple as a single transistor or op amp, with parts availability determining how low the cost.

What then can be improved in the circuit designed earlier? Well, today the TDA7000 single-chip FM receiver design still works well and is still in my shack functioning. However, several things have come to my attention concerning parts availability and cost in general. The TDA chip I used earlier is getting hard

to obtain, and the scrounging of component parts has created a delay factor for many amateurs in getting on 10 GHz quickly, not to mention 24 GHz, which is even more difficult.

Considering all this, I started to look at designing a new 30 MHz receiver (IF), but after looking at what was available on the commercial market, I have reconsidered designing a new receiver. What I discovered was a kit from Ramsey Electronics for a 10-meter FM receiver. This kit from Ramsey was very reasonably priced, making it attractive. Looking at Ramsey's literature and design specifications, I was quite taken by this

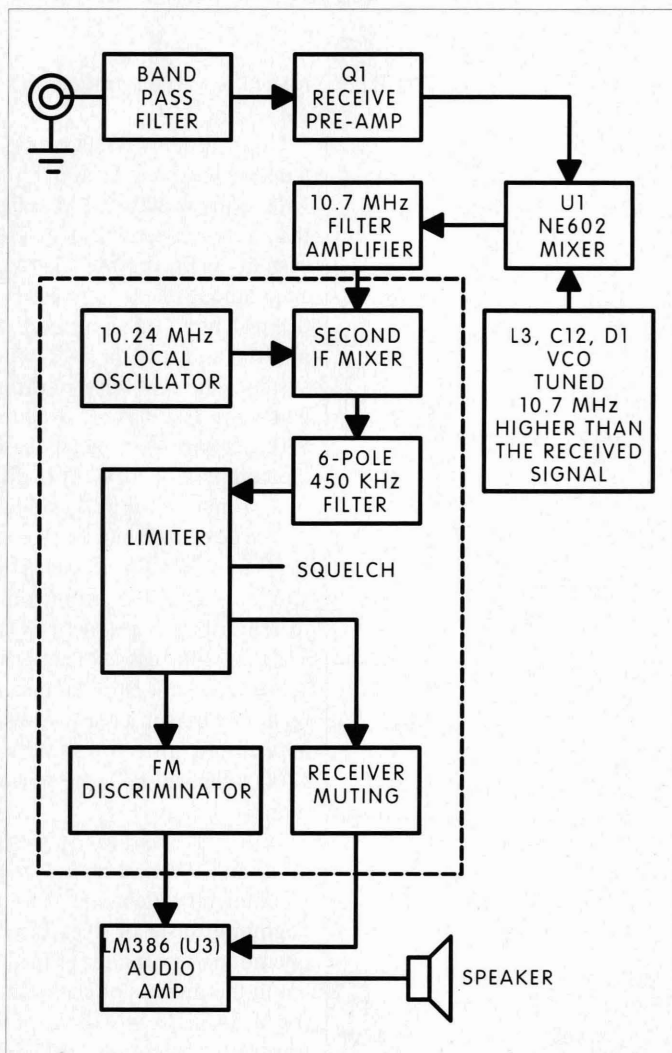


Fig. 2. Block diagram of Ramsey Electronics FR-10 30 MHz receiver kit with internal preamplifier, an excellent bargain that comes with all components needed to form an operational FM receiver. This receiver interfaces quite well with the Gunn diode transceiver system for either 10 GHz or 24 GHz operation requiring a 30 MHz wideband FM receiver system.

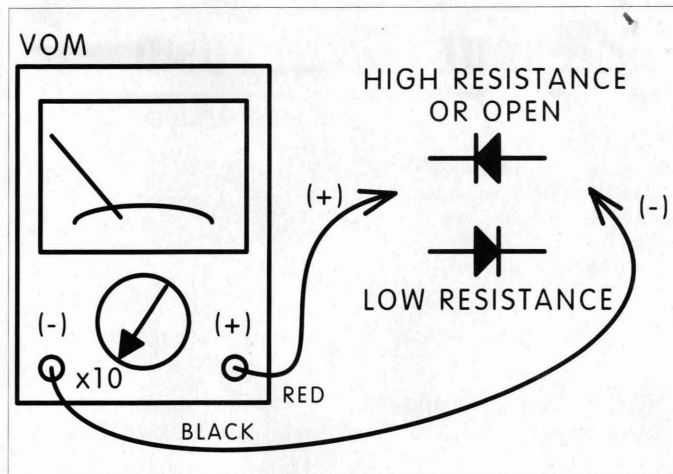


Fig. 3. Diode ohmmeter check to verify diode polarity. Application the same regardless of whether regular diode rectifier or varactor diode.

very fine and inexpensive FM receiver. The Ramsey FR-10 unit will interface well for the 30 MHz FM receiver portion of the Gunn transceiver with a few simple modifications to the original design.

Fig. 1 shows the functional block diagram of a basic wideband FM microwave system. This month, let's limit our scope to the 30 MHz wideband FM receiver, covering the assembly and modification of the Ramsey Electronics FR-10 receiver. Next time, I will cover the construction and assembly of the remainder of the system—the power supply modulator—and describe some testing methods.

Ramsey FR-10 FM receiver

I obtained two FR-10 receiver kits to construct two separate transceivers, one for Kerry N6IZW and one for myself, to allow testing of the finished product over a test range of about three miles. Performance testing is kind of difficult without a partner, be it on 10 or 24 GHz or somewhere else. It's kind of hard to talk to yourself, making a partner very important for experimentation on microwave.

The Ramsey kits arrived, and construction evaluation began right away as I was anxious to get started. Looking over the literature supplied with the kit, I was surprised with the detail

that was presented. It was quite sufficient to enable a beginner with limited construction ability to finish this kit and make it operational. The construction was presented in the Heathkit fashion familiar to many amateurs. An introduction is followed by a "how to construct your work area," a circuit description, and block diagrams on operation, all providing enough detail for familiarization with the FR-10.

After a review of the intro documentation, the component parts are checked. I might suggest a sorting procedure to avoid confusion when stuffing the PC board with components. Sort component parts onto pieces of paper with a pencil notation as to individual value. Place resistors on one paper so marked and capacitors on another. This makes for simple assembly without having to identify by color code or value once you begin construction. In sorting resistors, I always use a simple ohmmeter check to verify what these eyeglasses and magnifying lenses decode from the color bands.

A little preparation at the beginning might save you time later when you test the system out. Follow the construction details outlined in the Ramsey literature. The construction is quite easy, as each component part number is printed on top of

the circuit board to further ensure that you are placing the correct component in the proper location. The only error you could make is misidentifying and misvalidating resistor color codes or other parts values which are clearly marked.

I followed the instruction manual, and found that the pre-sorted components helped speed construction. All components were placed with ease—even polarity-sensitive capacitors, ICs, and transistors were plainly marked as to orientation on the top of the circuit board. Such a quality PC board makes errors avoidable if you take the time to read and look at parts identification on the board and in Ramsey's literature.

There was confusion regarding only one part in the kit—a varactor diode's correct orientation, the cathode/anode positioning on the board. While the board was clearly marked with which end was which, the diode was not. To verify which lead was the anode, I used my bench VOM meter. With most ohmmeters, you should use the x10 scale (current-limited) and the meter's red lead on what you think is the anode. That's positive side of the diode to the red lead and the cathode to the black lead (negative). The diode should indicate a few hundred ohms; if not, reverse the diode. When you get a reading of a few hundred ohms, you know which end is the cathode, as it is on the black lead of the ohmmeter in this simple test. Reversing the meter leads, there should be no indication on the ohmmeter.

To prove your test is accurate and that your meter is positive on the red lead, grab any marked diode, put the black lead on the cathode and the red on the anode, and verify a reading of a few hundred ohms. See **Fig. 3**.

This was no big deal. Only one part, and minor confusion solved by a simple ohmmeter check. Even if the diode were inserted backwards it would only have disabled the frequency tuning control. A simple

voltage check on the diode side of R8 would prove a fixed voltage of 0.6 volts when the diode was reversed on the board. When the diode is mounted correctly, the voltages will read from zero to six volts when the tuning pot is rotated.

A minor problem, but one that will not cause things to "go up in smoke" when power is applied. Give the PC board a good going over for solder bridges and clippings from component parts before proceeding. Make sure that the DC power input is free of shorts by using the ohmmeter, again on the x10 scale. Check the DC 9 V input leads and turn on the power switch. The ohmmeter should indicate low resistance and then rise to a much higher reading. What is changing is that the electrolytic capacitors are charging and the ohmmeter, in a few seconds, should be in the k Ω ranges. Now you're ready to apply power and check out the receiver.

I used my old URM-25 signal generator to verify operation and sensitivity at 30 MHz. Both receivers constructed performed as advertised, and sensitivity was less than a microvolt for full quieting. Follow Ramsey's instructions and all will be well. My two receivers worked right from the start. The only adjustment was L4 for audio output and L3 for proper frequency range adjustment. Ramsey thought of everything: They even provided a tuning tool to make the adjustments. What a kit! All I had to provide was my time and some solder!

Once I adjusted the RF frequency oscillator coil, I wanted to see how well the onboard RF preamplifier functioned. Was I surprised that the sensitivity was in the tenths of a microvolt for full quieting! There was some pickup of commercial FM broadcast in the 88 to 108 MHz range on very strong local stations. I had observed this before with other systems—even the TDA7000 circuit design, years ago. Note that at this stage of testing I have not placed the receiver and

modulator into a metal cabinet, which is necessary from a shielding standpoint. Any receiver that is going to be co-located on a hill-top with commercial high-power radio and TV transmitters can get into trouble if shielding is not used. If you use good shielding techniques, interference will be minimized.

I can't give more importance to shielding and to proper lead bypassing with feedthrough capacitors. All power leads that leave the cabinet should be tied through a feedthrough capacitor, not a rubber grommet. This goes for the hot speaker lead as well. With a metal cabinet and feedthrough capacitors used, the interference from commercial FM was eliminated.

Modifications for wideband FM

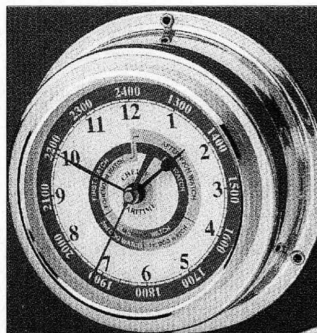
There are modifications that need to be made once you have determined that your receiver is operating properly. These modifications are necessary for greater bandwidth than the receiver comes stocked with. The conversion is simple and includes removing FL2 (450 kHz filter) and replacing it with a 0.01 μ F disc capacitor. This makes the passband about 100 kHz wide. Shunt resistor R14 (33 k Ω) with a 3.9 k on the bottom of the PC

board. This further broadens the passband to about 200 kHz.

The original tuning range of the FR-10 is about 5 MHz wide. We need to reduce this to about 500 kHz, centered at about 30 MHz. Remove C12 (56 pF) and replace it with a 15 pF disc capacitor. A second 15 pF disc capacitor is added under the PC board, shunting across L3, the oscillator coil.

To add finer frequency tuning, remove 10 k pot R10 and replace it with a 10-turn 10 k pot and calibrated readout dial for really fine frequency adjustment. Also, my headphones did not fit the headphone jack that was supplied with the kit, so I just jumpered the bottom of the board to a panel-mounted jack to fit my mike/headset connector.

The PC board was fitted with four standoffs to position it off the chassis by a small distance. The cabinet was connectorized with different connectors for each of the termination leads to avoid connecting the wrong device to the wrong jack. For instance, I used a two-connector cord for DC power, and bypassed the leads inside the cabinet with feedthrough capacitors. The headset has both a miniature one-eighth-inch connector and a standard quarter-inch stereo connector. This allows a



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Mobile T-hunting, international-style

"Won't you run out of topics?" That's what a former editor of *73 Magazine* asked when I proposed a monthly column on radio direction finding (RDF), better known as foxhunting. It's been over 10 years since then, and there are more potential topics than ever. If RDF is one of your ham radio interests, you can always find something new to do. There are projects to build, noise sources to eliminate, jammers and bootleggers to catch, wildlife to track, aircraft ELTs to locate, and hidden transmitters to find.

Recent "Homing In" columns have emphasized foxhunting under International Amateur Radio Union (IARU) rules. This on-foot sport, which usually takes place in large forests and parks, is called foxtailing, foxteering, ARDF and radio-orientteering. Many clubs in North

America are adding it to their hidden transmitter hunting activities. They are looking for ways to get T-hunters out of their cars and into the woods once in a while. One way to do that is to have mobile T-hunts that follow the foxtailing format, but with foxes spread out so that hunters can drive to them, at least most of the way.

I first heard about mobile versions of international-rules hunts from hams in Australia. (They call it CarDF.) Jay Hennigan WB6RDV and Dennis Schwendtner WB6OBB of Santa Barbara, California, were probably the first stateside hams to try it. According to the report from Marvin Johnston KE6HTS, everyone thought it was fun. So I decided to give it a try when it was my duty to put on the First Saturday Night T-hunt here in Fullerton last September. April Moell WA6OPS, Tom Curlee WB6UZZ, and David Curlee KE6IPY helped out.

Not enough transmissions

IARU foxtailing rules call for five transmitters on the same frequency, each transmitting a simple MCW message for one minute in sequence. Fox #1 sends MOE MOE MOE for 60 seconds; then it goes off and #2 sends MOI MOI MOI for 60 seconds. Then #3 sends MOS MOS MOS, and so on until #5 goes off and #1 immediately starts the sequence over again. That's the way my five Ts were programmed. Even without knowing Morse, it's easy to determine which T is on the air by counting dits.

In radio-orientteering, the goal is to walk or run to all foxes, then get to the finish line first and within the time limit, usually about two hours. Late arrivals are disqualified. For this First Saturday mobile hunt, which is traditionally scored by odometer mileage, we announced that the winner would be the team that had the lowest mileage among those finding the greatest number of Ts. Mileage would be taken at T5, so teams had to find it last. The others could be found in any order. All Ts were programmed to shut off at midnight, exactly five hours after the start of the hunt. Teams not reporting to T5 by then would be disqualified.

Traditional southern California mobile T-hunts are both a

"hider versus hunter" and "hunter versus hunter" game. The hiders do their best to foil the hunters using mountain bounces, not-on-the-map fire trails, camouflage, and so forth. But an international-style hunt is supposed to be a competition only among hunters, not between hiders and hunters. It's intended to teach hunters to take bearings carefully, plot them, plan an efficient route, and execute the plan.

In that spirit, we made every effort to have the hiding spots be relatively easy to find, with none of the usual signal tricks. All Ts were copyable at the starting point, though it was necessary to move around the start hilltop to get optimum signals on all of them. The fox boxes were right next to roads accessible by passenger car, except for two that were in public places less than 250 feet from roads. All of them were within 11 airline miles of the start.

The clear lesson from this hunt was that IARU's fox timing is not optimum for mobile T-hunting. The one-minute-on/four-minutes-off transmission ratio gives only 12 transmissions per hour per fox. Only one team found all five foxes, but arrived at the last fox 10 minutes after the midnight deadline. If we do it again, we will use a 12-seconds-on/48-seconds-off cycle for each fox, which will

choice between using a mini headset or my dual padded stereo headset for noisy locations. The mike jack is a standard three-pin miniature of different size than the receive headset jacks, avoiding confusion.

The modulator that provides DC power to the diode is fed via coaxial cable and appropriate connectors to easily tell the voltage supply from the detector (receiver diode). I used a BNC on the voltage connector and an SMA for the input to the Ramsey FR-10 30 MHz receiver. This further avoids confusion when in a

rush to get on the air. You cannot make a wrong connection ... unless you use a pipe wrench!

What is my recommendation on the Ramsey Electronics FR-10 30 MHz receiver kit? If you are interested in 10 GHz or even 24 GHz wideband operation, this is a most inexpensive kit that performs even better than I expected. You cannot ask for better value in a kit. Ramsey even tossed in the alignment tool, to be sure you could adjust the two coils in the kit properly. That in itself shows that Ramsey

should be proud of the kits they produce. The FR-10 comes not only with a quality PC board, but also with tools and easy-to-follow instructions to ensure that the kit will function as advertised.

I constructed two kits. Both were simple to assemble and both functioned better than I expected. Sensitivity was a key issue and the 0.2 microvolt sensitivity for full quieting of the receiver at 30 MHz was very impressive. Order your FR-10 receiver kit now and be ready for the completion of this

project in Part 2 next time. Ramsey Electronics is located at 793 Canning Parkway, Victor NY 14564. Their order phone is 1 (800) 446-2295, while questions or order status handling is available at (716) 924-4555.

Next time, I plan to complete this transceiver and cover the interface of the receiver and the transmitter modulator circuit. The bottom line is that this receiver can be interfaced with either a 10 GHz or 24 GHz oscillator/detector cavity for great wideband FM fun. 73, Chuck WB6IGP.

ABOVE & BEYOND

VHF and Above Operation

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Shakedown tests of portable microwave systems

August is the month for the start of the ARRL 10GHz Contest. It is open to all participants who have access to or have constructed transceivers for making contacts on 10GHz. The rules are quite simple—communicate with one another and promote activity and experimentation on our microwave bands. Well, there are a few more rules to follow, but the main spirit of the contest is to encourage communications among amateurs and to promote 10GHz microwave communication.

Most amateurs participating in this contest operate either wideband FM or single sideband. Most of the old-timers on this band have migrated to narrowband operation for its superior performance with single sideband operation. Even so, many still take their wideband rigs along just to keep the simple rigs in operation and to

make contacts with ever-present new amateurs.

I am sure that in the future (if not in this contest, then in another) there will be amateurs who might attempt high-speed data communications. Some others may use standard video for communications. Art WA6YQH and Ray N6RE could be ready to use their 10GHz facilities for making several contacts using full-scan NTSC video. They have been the video pioneers in our group, making some very important contributions on how to set up systems and methods for video operation on 10GHz.

The beauty of operating on 10GHz is that there is space for all of us to improve our own skills and construction practices. On 10GHz, there are sufficient frequency allocations allowing all these modes of operation.

In our very first years (early 1980s), I operated on wideband FM, as this was the only equipment that was (1) inexpensive and (2) available (inexpensive) at that time. Now, I utilize an improved system that is fully linear in operation. It will pass FM as well as SSB signals equally well. We even tried

video operation briefly. My main mode of operation is SSB—distance and weak signal work make this mode of operation the best choice.

When signals are moderate to very strong, we usually switch to narrowband FM. The 5kHz bandpass with FM allows for very crystal clear communications without retuning, as is the case when you work many different stations using SSB. Operation on FM is not unlike using a standard 2m HT, as quality is quite the same.

The ARRL 10GHz Contest is my cup of tea—one that lets us commiserate while participating in something we all enjoy. The annual contest is usually held on the third weekends of August and September. Get a copy of *QST* for contest rules.

Sometimes preparation for the contest is put off till the last moment, and that's what this column is all about. First, to get you thinking about getting in on the fun on 10GHz after you have put together a 10GHz transverter, I want to provide a list of functional tests to ensure that it will work after you transport it to a remote hilltop.

Most things we purchase are expected to operate in almost every environment, whether fixed station or portable/mobile. However, things we home-brew tend to need a little more care, depending on the packaging and construction methods. In this arena, I cast myself into the caldron as I have failed to do, several times, what I am about to tell you to do. This is the voice of experience and near failure speaking: If you do not test prior to going to a remote location, you just might not have a functional rig, or at least a very sensitive one, to use. I have found out on more than one occasion that a moment of preparation prior to going to the hilltop will be well-spent time.

What you need to do is to check out your transverter and make sure that all is well. Here the best test is to get several of your buddies together and test all the rigs in your group. You might not have the latest in noise figure or frequency test

equipment—this is not needed but nice to have. What you want to find out is how you fare compared to the group of fellow amateurs with whom you expect to communicate.

Is your antenna properly focused, and are your signals being received in similar strength, as well as transmitter power? You're not after an absolute value, but a simple comparison of systems, noting the top performers and finding out what is out of adjustment or not sensitive enough. In general, a full system shakedown performance/repair/adjustment/evaluation on home grounds.

Checking out your rig sounds simple even for a well-equipped home shack, as far as test equipment goes. Remember, even if all seems well, just making the trip to some remote spot with your equipment is the same as placing it on a "shake table." For me, this shake table is the back of my pickup truck. Even though the equipment was carried in boxes, it was still not transported in a "hardened" or shock-resistant state for an arduous trip to a microwave hilltop site.

My transceiver for many years has been a home-constructed system fabricated on the top plate of a 10GHz, 10W TWT amplifier. There are two different DC-to-DC power supplies providing -24 volts, +12 volts, -5 volts, and +24 volts. All are isolated switch mode power supplies allowing a single common ground for all components between input and output of the supplies. There are a microwave brick-type Frequency West oscillator, a mixer, and two 10GHz amplifiers, all connected to a bank of four SPDT microwave relays. In addition, there is a circuit that will detect RF on the IF cable between the transverter and 2m driving radio.

Whatever your system, place it in an environmentally suitable antenna test range in competition with others for some friendly tests. See who has the most sensitive receiver and best power-to-antenna performance. All things are equal when you factor in power output antenna gains expected. By observing

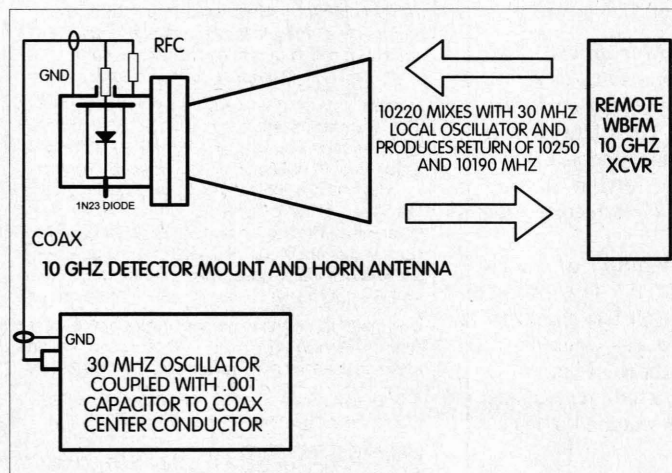


Fig. 1. Basic microwave "boomerang." This device will provide a return signal for a selected band of operation. It is the simplest one that can be constructed for test purposes using wideband FM-type Gunn-diode oscillator receivers/transmitters. Uses local oscillator running at IF frequency of 30MHz injected into waveguide transition and small horn antenna.

the remote system signal (IF) sensitivity returned in either S-units or so much power as observed on a sensitive RF power meter, you will get some very quick ballpark figures of some really smoking signals and some other performers that should be as good. This test points to a system performance level and may show that some part of the system is not functioning as well as it should be.

The problem could be as simple as changing the feed position on a dish antenna to illuminate it better than what you are getting from where the geometry says the feed should be. Small adjustments can be best made in this type of setting—this is how the pros do it and certify their equipment. Why not take advantage of their antenna range adjustments and apply it ourselves? True, the backyard antenna range is simple, but it's still effective in determining just how well your rig stacks up against others. See **Figs. 2** and **3** for information on how we set up our test range.

The remote transceiver is the most basic unit that can be constructed for the frequency band in question. In our case, the transceiver is constructed for 10GHz operation. It is built with a Frequency West microwave brick-type oscillator operating at 10,224MHz injecting into the LO (local oscillator) port of a microwave coaxial mixer. The

mixer output (RF) port is coupled directly to a small waveguide horn antenna that has about 10 to 15dB gain. The horn gain is not important and should be kept low, as we want a weak signal to detect at the far end of the test path.

If signals are too strong, a waveguide or coaxial attenuator can be placed in the connection between the mixer and the antenna. The IF port of the mixer is connected to a long run of coax back to the point of test. In our case, a 300-foot length of RG-59 was used to allow insertion of amplifier pads, power meters, and signal generators to make performance evaluations.

By using a signal generator for the transmitting source, output power can be controlled by the quality attenuator on the generator, sending a measured amount of signal to the remote transceiver and seeing how much is required to detect a minimum discernible signal on the test receiver system. The main reason we do not want to use a 2m transceiver for these tests is that the S-meter is limiter-controlled and can give bad information when the signal is in saturation. Using the signal generator with variable power output, it can be set to a calibrated point and referenced to detected sensitivity. These tests are not absolute, but in a practical sense are very good.

It doesn't matter what you choose to use in your system, be

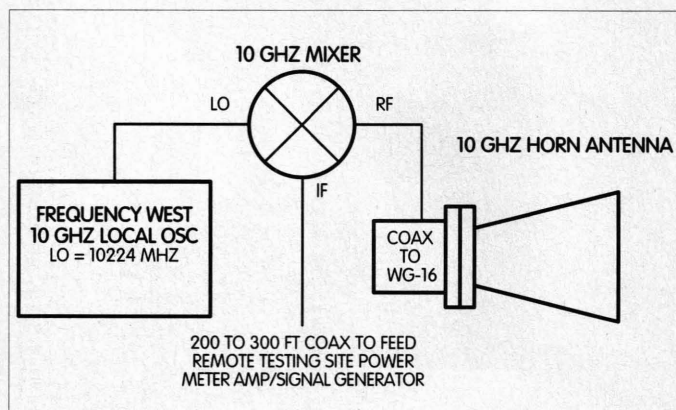


Fig. 2. Sensitive simple transceiver consisting of local oscillator and mixer fixed to small horn antenna. Operation is linear, so FM, SSB, and video can be run on system. The transceiver is mounted in metal housing to protect it from the elements and provide orientation, horizontal and vertical adjustment for the remote end of the antenna test range. In operation it's a very low power test receiver/transmitter.

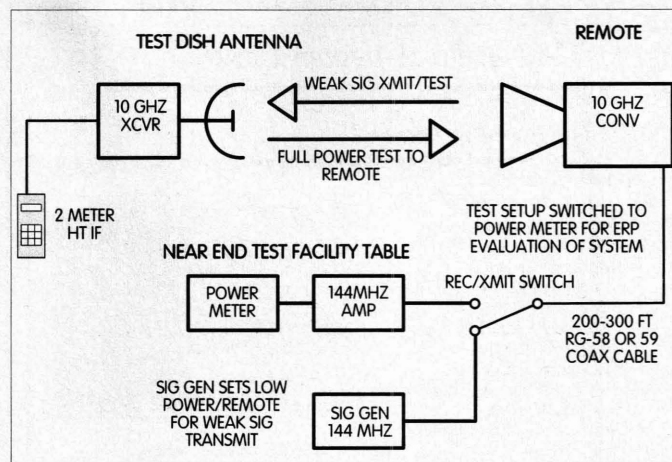


Fig. 3. Test-range setup using a testing path constructed in Kerry N6IZW's backyard, with a remote flea power transceiver as shown in **Fig. 2**. The unit is placed at the far end of the backyard—about 200 feet away. The coax at IF frequency of the remote unit (144MHz) is returned to near end of antenna test range for system evaluation. This provides a relative evaluation of system sensitivity and transmitter output power, including evaluation of antenna performance.

it expensive signal generators and power meters or a 2m handie-talkie. Just be ready to adjust your system, and at minimum with the 2m HT have a variable attenuator or series of coaxial pads (fixed attenuators) to reduce its power output as well as sensitivity on receive.

Run the 2m HT on its lowest power setting. I am assuming 1/4W or so (that's 250 milliwatts), or +24dBm. Remember here that the mixer only needs a maximum of +10dBm for full function.

Adjust your 2m power to conform to these limits at a maximum or else you will just "let the smoke out of" the mixer—and it can't be put back. The loss of a microwave mixer is indeed a sad occurrence. Just be careful and keep power to the mixer in transmit to a low level. When setting up the test range, measure the loss of your long run of coax and include it in the formula you develop for your situation.

Did anyone catch that I stated I used RG-59? Yes, it's not 50Ω coax but 75Ω, and the SWR will not be a large factor in the tests because the coax is so long the effect is quite minimal. Why did I use RG-59? Well, primarily because it was the only coax around that was 300 feet long and in one piece. I did not want to put together a nightmare of splices. Due to the length used,

it was not found to be a problem because we needed all the additional loss we could find to make reasonably meaningful sensitivity tests.

How did my rig fare in these test evaluations? Well, I don't expect to take the cake in the receiver sensitivity tests as I still use a pair of MGF-1402 GaAsFET front-end preamplifiers whose noise figure is in the 2 to 3dB range. There will be more sensitive receivers in the test pool. I just haven't had time to take my own advice and improve my receiver noise figure by replacing the MGF preamps with a better GaAsFET at 10GHz. Nonetheless, a transmit power output of 10W and my 30-inch dish should perform on a par with the other systems being evaluated.

Now off to the trials to find out if what I expect to see will be true. Our test is scheduled for an evening meeting of our microwave group, so I hope we get going while I can still take some photos!

And the winner is ...

Well, we had the test-range evaluation, and as I suspected the receiver sensitivity of my 10GHz front end was able to detect a signal as transmitted at -80dB into the test system. The better systems with lower noise figures were able to make minimum

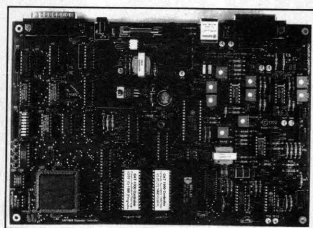
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Photo A. Photo of N6IZW's backyard, showing rigs to be tested. Left, Ed W6OYJ; at table, Kerry N6IZW; John WB6BKR; author's rig; and Jay KD6PBH. Test-range test performed by Kerry N6IZW at test table.

discernible signal detections to -85dB.

My system was working well but not top-notch here. The receive results show that I need to improve the signal-to-noise ratio by replacing the first preamp with one of a lower noise figure. As far as transmit goes, we calculated that my system (looking at dish diameter gain and power output of the TWT amplifier at 10W) gave an ERP of 73dB. This compared favorably with Kerry N6IZW's system. Both of our systems use the same type and power TWT amplifier

and dish antenna so they should be quite identical in operation.

Next month: the world of coaxial switches. I plan to get into specifics so that you can recognize what makes a good VHF or HF switch and what the differences are. Of course, these switches can be found in surplus as we try to hold down costs. The main thrust will be to show the different types that are available in surplus and what frequencies they are suited for. If you have any questions on this or other topics, please write. 73, Chuck WB6IGP.

73

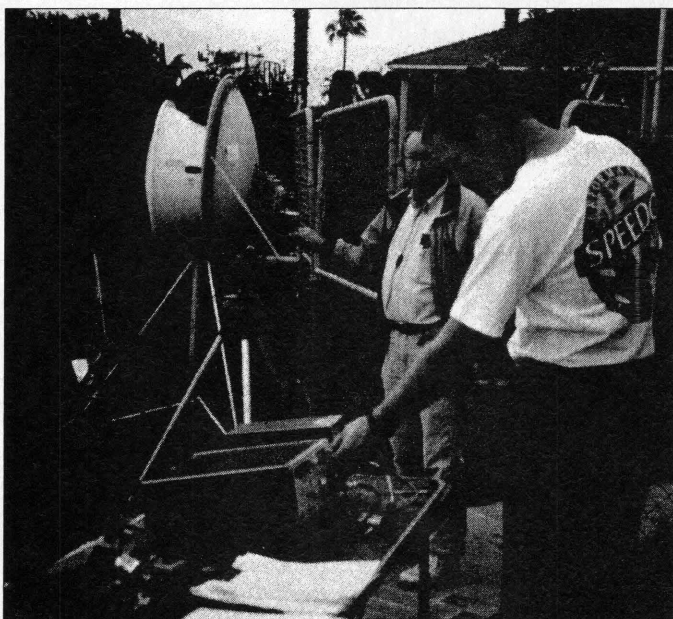


Photo B. Close-up of test table and N6IZW checking receiver sensitivity of Ed W6OYJ's 10GHz receiver. What looks like one odd-shaped dish is really two dish setups brought for test (see Photo A).

VHF and Above Operation

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Making antenna measurements

This month I want to describe some experiments with microwave and VHF test equipment and antennas that were covered at one of our San Diego Microwave Group meetings at Kerry N6IZW's home. These experiments covered some of the most basic of all antenna principles, demonstrating them on small models. The ability to see on a small-scale equivalent proved invaluable; it clarified our understanding of antennas and effects that demonstrate antenna principles.

We all are aware of these basic principles, like put the antenna in the clear at least one-quarter wavelength above ground, place it away from obstructions in the transmission path, and on and on. I, too, have been pretty much aware of what I have read, but due to the sheer size of antennas and their supporting structures, I don't have the space/time/inclination to test these principles. I just put up the best antenna and structure that I can

manage and let it go at that point.

The antenna test range

A short antenna demonstration range was set up on a tabletop showing how RF propagation is affected, and some principles applied to its path losses from source (transmitter) to destination (receiver). 10 GHz microwave frequencies were used in this demonstration, which allowed a tabletop to function as the test range. The table, being about five feet long, provided a test range nearly 70 wavelengths long, making a good evaluation on a simple basis. 300 million divided by a frequency of 10250 MHz equals .29 meters or 3 cm (1.14 inches) for a wavelength at 10250 MHz. See Fig. 1.

Does this make any sense when you think about a 10 GHz waveguide (wavelength being one inch and waveguide for 10 GHz being one-half by one inch in physical size)? With a wavelength being about an inch and the table being five feet long,

each inch of that separation is equivalent to a full wavelength of RF radiation between antennas. Try that at 40 meters—if you do, you'll need the entire farm and *that* might not be enough.

The tabletop test

The tabletop antenna test range was set up to have a microwave power meter connected to a coaxial waveguide transition and a small horn antenna at the head end of the table. (The transition used was very similar to the transition described in last month's column.) At the opposite end of the table was a Gunn diode oscillator, equipped with similar horn antennas as the power meter end of the path being tested.

The power meter (destination) and the transmitter (source) were tested to each other (without a table in the path); there was little difference noted in the power meter reading as the transmitter source was raised to a higher position of alignment than each horn antenna. Originally the horn antennas were pointing at each other at similar heights. Well, you say, what does this test show? This test proves that for two antennas aimed at each other, with the transmit antenna at varied heights, there will be no change in the receive antenna signal strength. If the path changes with height you have secondary path interference.

If there was a secondary reflection path it would arrive with phase difference, in reference to the main signal path. I refer to either the addition of energy or the subtraction of energy, due to the signals being in or out of phase with each other. A simple test to set up is to place something under the antennas that would reflect energy. This test simulates the earth and possibly a body of water or structures in the path between antennas.

With the table back in place, space the antennas about a foot above the table surface (exactly the same position as the

previous test, without table). Place a piece of tin foil covering the length of the table between the antennas and repeat the same test as before. The tin foil will simulate the earth, structures and bodies of water. Slowly increase or decrease the vertical position of the transmit antenna and source, while watching the power meter for any change in dB level. Note as you change vertical height there is quite a change in power meter reading. There will be a point where you can reach minimum and maximum meter indications.

This change that you now see is directly affected by the tin foil reflecting part of the microwave signal into the detector with the main path signal. This is ground reflection, or multipath, affecting the total path performance of the microwave signal. It's kind of a slick trick to see a complete antenna test range on top of a table. I find it very interesting to demonstrate in miniature the test range and see firsthand things we have read about antenna structures, but because of the lower frequencies involved, we were not able to create these types of effects. You can read about them but the demonstration brings the message home much better.

Polarization tests

Another experiment can be one test of polarization, that is, where both antennas are vertical or horizontal. Is there a penalty in the transmission and reception if one antenna is vertical and the other horizontal in polarization? This test can be made with the test range from the last experiment—just remove the tin foil from the test setup and note the power meter reading. Let's say it was reading +5 dBm for a point of reference. The number is not important: Just record what you are reading, for reference in dB power.

We're assuming that the transmitter is mounted on a small camera tripod for these

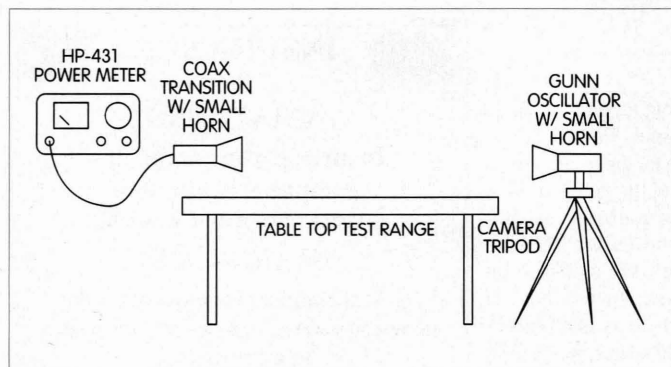


Fig. 1. Antenna test range constructed on a tabletop. The microwave power meter is located on one end and the test transmitter (10 GHz Gunn oscillator located on small camera tripod) is on the other end of the test range.

tests. Unlock the swivel mount of the camera tripod, and while noting the power meter reading, slowly rotate the mount into a vertical position. Notice the power meter as you rotate the antenna from one polarization to another. If things are proper you should see something near 30 dB, all things being equal. We saw a dB change on our test range of 26 dB, as read on the power meter. This was attributed to our test fixture and influence from reflections in the garage. For demonstration purposes it worked well.

Circular polarization

What other tests can you perform in such an antenna test range? Let's try circular polarization. This is neither vertical nor horizontal but can be described as using a spiral type of antenna when incoming signals are of various polarizations. You are familiar with crossed polarization antennas that are used for satellite communications. They differ from circular antennas in that they can be switched to either vertical or horizontal by a relay at the antenna.

As the name implies they are cross-polarized, in that two single-beam antennas are combined on one boom to support both polarizations, vertical and horizontal. Usually for gain purposes two sets of such antennas are mounted horizontally in reference to the antenna tower or mast.

The circular antenna can take several forms, one being constructed like a coil of wire as it is taken off a spool and allowed to drop to the ground in coils much like a stretched Slinky™ toy. This type of antenna is frequency-dependent on a center design frequency. It is not sharp, but rather broadbanded in nature. See Fig. 2 for circular antenna details.

Another form of spiral, or helix, antenna can take the shape of a modified coil, or helix, in that the coils start at the feed with a very tight coil and as the length of the antenna increases

so does the spiral increase. It's much like winding a spaced coil of wire on a pointy ice cream cone.

Still another version, used for condensed space, is a small flat circular spiral etched on a piece of PC board. The spiral is very tightly wound of really closely-spaced printed lines etched on the circuit board. This antenna is broadband in frequency range from several GHz to about 12 GHz. This type of antenna is primarily for radar threat reception and similar uses.

These antennas are usually found in surplus, looking much like a can of tuna with a coax connector on one side, and a plastic or epoxy lens on the front side. Some show a spiral pattern on the plastic side through the plastic cover. Other forms could be of similar diameter but only a quarter of an inch deep, with a coax connector, again, on the back side of the antenna.

In any case, replace the horn antennas on both ends of the tabletop test range with the spiral circularly polarized test antennas. Repeat the previous test as to vertical/horizontal movement of the test antenna. Note the power meter reading this time, when the test antenna is rotated between vertical and horizontal patterns. Did the meter change in level from the test rotation?

Repeat the test on multipath with the spiral feeds. Replace the tin foil on top of the table before running this test. We did not have time to finish this test at our monthly microwave meeting. Give both these tests a try. Just had to leave you with something to ponder and speculate on when the spiral is varied in height above the tabletop. I will answer that question next month.

If you want to work out a pattern of the antenna you have under test there is a simple platform to construct that will manually let you test an antenna, be it a horn spiral or other 10 GHz antenna on this test range.

What you need to do to evaluate the antenna pattern is to place the transmitter with its antenna to be tested on a calibrated turntable. What you should do is to scribe 360-degree compass calibration marks on the turntable from an old record player turntable and rotate the entire antenna and turntable while taking readings.

If you're diligent enough, you can make checks every degree or so, to evaluate what the power meter sees from the test antenna, as you rotate the test structure turntable and antenna about the compass headings. The best way to do this test is to take the test antenna range out of the garage and out into the open to avoid reflections of nearby objects interfering with the transmission of energy between the two antennas.

Commercial manufacturers have automatic equipment to perform tests and create examples, but we can do similar,

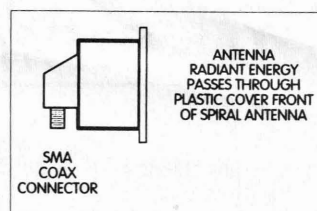


Fig. 2. Several types of circularly polarized microwave broadband antennas are similar; shown is an example of a typical military surplus model.

if less high-tech, testing. Sure, it takes time to make a lot of readings, but when you plot them out on a piece of graph paper you will have a picture of what your antenna's beam pattern looks like from this simple antenna test range.

I cannot state this more strongly: Recording many readings like this gives a true picture of just what an antenna is doing—be it at 10 GHz, or at lower frequencies where you cannot repeat the test, because

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CIRCLE 13 ON READER SERVICE CARD

THE DIGITAL PORT

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You can still build your own packet modem, and this may be a fascinating project for your club. In past columns I have covered the ease of building a serial modem for the HF digital modes and lamented on the lack of the all-important TCM 3105 to build a VHF packet modem. Through the thoughtfulness of Dave WA4GVT, I found LDG Electronics still has a supply of ready-to-go kits to build your own low-cost 1200b packet modem.

I don't know if there will be more kits after these are gone, but there are probably enough for those who wish to take advantage of a most reasonable entry into packet radio. The price tag, at \$30, should be enough incentive to grab the interest of even the most frugal prospective packeteer.

The cost will cover all you need except for connectors, cables and an enclosure. You will find instructions to build your own cables are included, so you can save a little there. And it will perform, of course, without the enclosure. There is a terminal software package included, plus there are a lot of programs available elsewhere to

make this project successful and fun to experiment with.

The disk in the kit contains, along with the program, a comprehensive educational manual that explains packet operation, plus explicit instructions for assembly. The printed circuit board is plainly marked, and by following the step-by-step instruction sheet, the assembly is simple and straightforward. All in all, it's an enjoyable experience.

You will also find a parts list that gives Radio Shack™ numbers for connectors, cable parts and a suggested enclosure. I found the box was no longer available and substituted a part # 270-1802. This worked well, though it was necessary to elongate the mounting holes on the PCB slightly to fit the bosses in the box.

Also, you will want to avoid one of my forays into near futility. Since I wanted to use the same 25-pin cable to this modem as I use on the HF modem and the BayCom™, as well as the PK232, I installed a connector on the end of the box that just does *not* belong. It will survive and doesn't look really awful, but only after drilling and tapping, epoxy, plus modifying

the scale of materials is too large. That's the beauty of using microwave as a modeling structure, to help explain what is happening and how the antenna is performing.

The tests you make and the data you obtain will be usable not just at 10 GHz, but will also be applicable to other frequencies. It's like working with a model train set on the living

room carpet. While we know it's just a scaled-down version of a real train, its operation is just the same, be it 100 tons, or just a pound or two. The difference here is size, and with the antenna test range, it's still size—and scaling down to test for similar features and effects. Modeling allows you to demonstrate antenna performance on top of a table. 

the connector. A nine-pin connector is the best advice. Of course, a plastic shoe box would offer more leeway for the untamed builder.

It works great

After completing the hardware and cabling stage, I brought up my copy of BayCom 1.4 software. It has been the most reliable of all I have tried for use with the serial modem, so I like to rule out failure at first, then advance into unknown waters later. The Basic Packet Modem is a sparkling success. It connects to the local PBBS with speed rivaling the PK232.

I found the previous source of this version of BayCom software is no longer at the Web site where I first discovered it. After running an Internet search, I could only find Version 1.5 (I've listed the address in **Table 1**). It is a DOS program and the documentation that comes in the separate "Manual.zip" will lead you through the setup and operation.

As with any terminal program, the configuration must be correct. So ... print the manual, read it and follow the instructions. It is an education in itself.

As I read and listen to others, I hear similar praises for this program. I did find Version 1.4 in other than English versions, but even those locations are getting sparse. Version 1.6 is out and not a freebie. I have not had my hands on it, but presumably the commercial version is a tad more user friendly. Take this as your clue that the freeware version takes a little getting used to.

The often mentioned freeware from George SV2AGW is listed in **Table 1**. George is continuously upgrading his programs and they work very well. I found, in this instance, that my local node was not responding as well to the commands from this software as I would have liked. You will find performance varies from node to node, so experimentation is a key here.

Incidentally, George tells me he has been working a gateway

onto the Internet in Europe with an interface he wrote and surfing the World Wide Web with Netscape™ and similar browsers. There is a glaring difference between the possibilities as we see them in the States and European packet radio where they are running at 9600 baud. I hear numerous complaints that we need to upgrade from the snail-like 1200b. ... Someday.

Speaking of other software, I received a few messages from hams that made me rethink my position on PCFlexnet. Don WL7NF sent me info via packet that there is a driver on the PCFlexnet Web site that will work with the SoundBlaster™ cards. I had stated previously there was no such module. Just to be sure, I received word from Jim AB8AB that he had the system up and working with just a little tweaking left. So ... it can be done. Just leave it to the persistent ham to prove it.

A recent visit from Hal KC7STU made me painfully aware of trends in ham radio, and that some of this digital technology that we either ignore or take as a matter of course holds a large degree of the answer to breathe new life into this hobby. Hal is a software engineer and the things he saw that day obviously impressed him.

Some of this is not exactly new technology. Audio Frequency Shift Keying (AFSK) technology has been around for a few decades. Many of us ignored it, preferring the "real modes" of phone and CW. Besides, the cost of using digital modes was once prohibitive. You could scrimp and save to buy a radio that got you on the air. Why double the cost for something that tied you to a keyboard?

It has all changed

Hal could see there is something more to ham radio than a two-meter handheld and repeaters—and this was something he could afford and for which he could perhaps write some software to satisfy his needs. After

VHF and Above Operation

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10 GHz fun, 1999 update, part 2: the Gunn diode modulator power supply

Last time, we covered construction of the Ramsey FR-10 30 MHz FM IF system for our microwave transceivers for use on both 10 GHz and 24 GHz. This month, I want to complete the construction of the transceiver package, with discussion about the additional circuitry required in the transmitter portion of the system.

The power supply modulator in a wideband FM system is quite simple in that DC voltage is used to power a Gunn diode in a microwave cavity. The 24 GHz Gunn-varactor-controlled transceiver can be obtained from SHF Microwave Supply [arutz@shfmicro.com]; phone: (123) 456-789; FAX: (123) 456-789. The 30 MHz receiver was obtained from Ramsey Electronics.

793 Canning Parkway, Victor
NY 14564; 1 (800) 446-2295
will get you the order desk for
the FR-10 30 MHz receiver.

A little review is in order due to differences between 10 and 24 GHz Gunn oscillators. For 10 GHz, the Gunn voltage is in the 5 to 10 V range. Current requirements depend on the power output of the Gunn device. Ten milliwatt Gunn sources draw about 50 to 100 mA of current, while 100 mW devices can draw as much as 600 mA. 24 GHz Gunn diodes require lower voltages to function than the 10 GHz devices do. Nominal voltage for a 24 GHz Gunn device is in the 3 to 6 V range, with requirements similar to those of the 10 GHz devices with regard to power and current drawn.

The power supply/modulator for either circuit is quite the same. In each case, the power supply is constructed from a

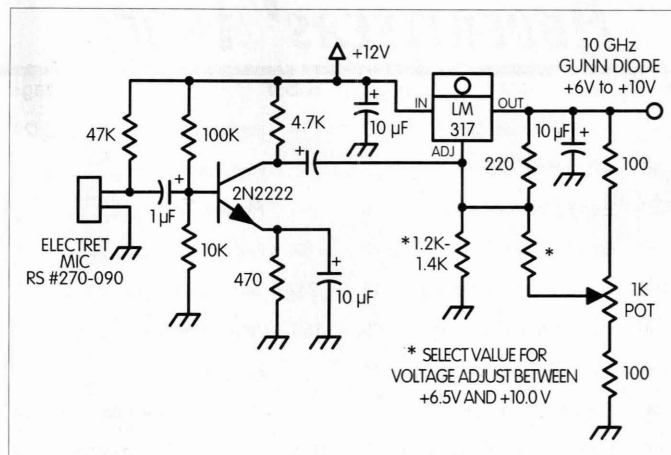


Fig. 1. Schematic for power supply modulator for 10 GHz Gunn diode source requiring +10 volts without varactor control. You must use an LM317 adjustable regulator for the circuit to function with modulation. A 7810 voltage regulator will not function as a regulator, having only in/out and ground, and no reference terminal.

single LM317 adjustable voltage regulator. For systems that use a varactor diode, the Gunn diode voltage is set at a fixed value near its maximum voltage of around +5 V, depending on diode specifications for that particular diode. Then, to adjust frequency, another variable resistor varies voltage on the varactor to adjust frequency of operation.

The modulator mike amplifier of the circuit can be a single transistor or an op amp. In the case of varactor cavities, the mike amp is connected to the

adjust terminal of the varactor regulator. Audio from the mike is a small-value AC component now riding on the regulator adjust terminal of the variable voltage regulator.

When the mike audio (a small-value AC voltage) is added to the fixed DC voltage on the regulator, it causes the output voltage to vary at the audio rate, producing a change in frequency varying at the audio rate. This produces FM (frequency modulation) on the transmit signal. The amount of

The Digital Port

continued from page 47

their kits. They were the ones who supplied the hard-to-get packet serial modem kit that I wrote about some months back and they have a reasonably priced tuner in kit form, or it can be purchased assembled. Their Web site is also listed in **Table 1**.

I see that the packet serial modem has been removed from their new Web site. It could be (though I doubt it seriously) that when I wrote about this wonderful packet modem that you folks simply cleaned them out and that was the last of the inventory. More likely, it was such a small item and was a bit

temperamental and possibly required more technical service replies than the profit could cover. The market has passed away on this item. There are a number of reasons as I discussed in a previous column.

Anyway, there are eight items listed and you might want to take a look. In addition to the regular tuner I intend to order there is a low-current-draw QRP tuner, two small computer boards for special control projects, relays to work with the computers, a repeater voting system, a balun kit and a nifty-looking headphone/speaker box to use between the output of your radio and your computer speakers.

One more item that looks like a winner comes from Timewave,

the folks who absorbed AEA. They already build a whale of a great DSP unit, the DSP-599zx (which is a must-install for the mobile installation here), that works very well when coupled with the old iron-horse AEA PK-232, and a great RTTY program to use directly with the modem in the DSP-599zx.

Now they have developed something unique just for the PK-232. They offer a DSP upgrade that works in all modes, including RTTY, PACTOR, CW and packet. This makes it possible to have excellent DSP performance for the PK-232 for 125 or 150 dollars depending on whether there is already a daughterboard in place. You can read about it on their Web site (see **Table 1**).

I had a recent E-mail asking where to purchase a PK-232. I replied that I had seen a number of used ones on the market in the \$100 range. I might also mention that if you look quickly, there may still be a closeout special from Timewave on the DSP-232 Multimode for \$100 listed on their Web site. I have had my PK-232MBX for so long that I wouldn't consider trading it, but that new \$100 multimode would sound good to someone who has none.

If you have questions or comments about this column, E-mail me at [jheller@sierra.net] and/or CompuServe [72130,1352]. I will gladly share what I know or find a resource for you. For now, 73, Jack KB7NO. **73**

audio when increased affects the deviation of the transmitter FM. The audio voltage (very small AC voltage) rides on top of the DC varactor control voltage that is used to set the RF frequency of operation.

Quite a simple scheme, mike audio to FM in a voltage regulator circuit. In the Gunn diode without varactor frequency control, the audio is connected to the Gunn diode voltage regulator adjust terminal to function much the same as in the varactor scenario. The non-varactor cavity setup is hampered with less frequency agility than varactor cavities allow. Frequency agility is quite good with varactor cavities, making them more expensive and desirable. In any event, both work—it's just that the varactor cavity is like a Lincoln in comparison with an economy car. See Fig. 1 for the power supply modulator circuit for a basic Gunn oscillator cavity without varactor control.

A simple circuit uses a single 2N2222 NPN transistor for the audio amplifier mike amp, as shown in Fig. 2. The LM317 circuit is similar for all applications, whether with 10 or 24 GHz Gunn sources. The only differing factor is the voltage required for the Gunn diode—

approximately 10 volts for a 10 GHz diode and a value of about five volts for a 24 GHz diode.

With most systems operating from +12 volts DC, a direct connection to the LM317 will be sufficient, with a modest heat sink to dissipate heat. For higher-current operation for high-current diodes, use a bootstrapping NPN pass transistor to increase the regulator's current handling ability. Almost any NPN transistor will work. I used a TO-220-case 2N3055, as it was in my junk box. Any modest current device will work, too. Use an insulating mount to secure the transistor to a chassis, as the back of the device is the collector and needs to be insulated on the heat sink. See Fig. 4 for circuit details.

Bypass the emitter of the NPN pass transistor with a 10 μ F or more cap (value not critical) to minimize noise on the DC line from the regulator. By looking on a scope, I found that at this emitter output point I had quite an AC oscillation when the regulator was combined. I eliminated the oscillation with a 40 μ F capacitor between the emitter of the 2N3055 and ground. I just grabbed the first tantalum out of the junk box—I suspect that a 10 μ F would work just as well.

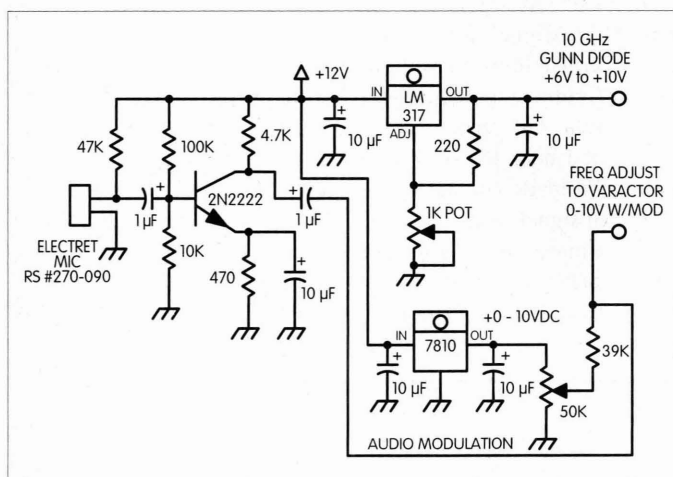


Fig. 2. 10 GHz schematic changes for varactor control mike audio applied to varactor for FM modulation. Audio voltage is superimposed on top of varactor DC control voltage that is used for frequency tuning. The LM317 could be replaced here by a fixed 7810 voltage regulator and used for both Gunn and varactor supply. Two regulators are used in this example for demonstration purposes.

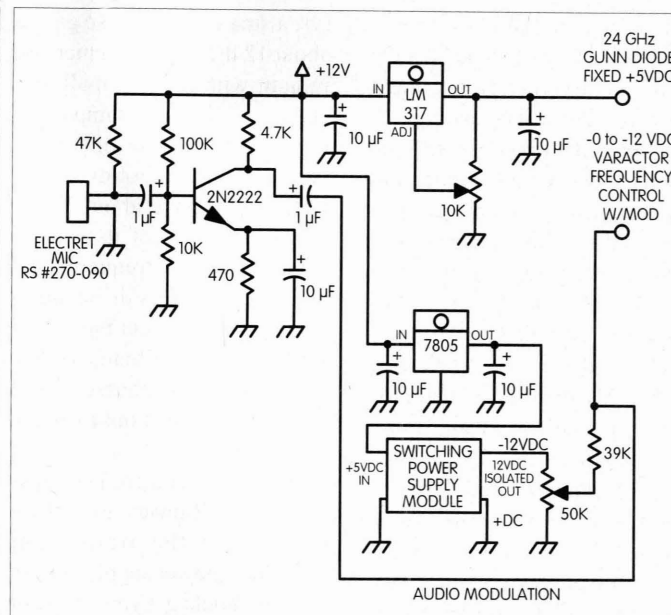


Fig. 3. Schematic changes when using 24 GHz Gunn diode source and varactor tuning arrangements. Note that on the 24 GHz cavity the varactor uses a negative voltage for control of frequency adjustment. Gunn diode voltage must be reduced to the required 5 to 6 V range. Verify your diode's maximum voltage before applying power. Note the addition of a small isolated switching PC-board-mount power supply added to obtain the inverted negative output for varactor tuning voltage. The power supply can be very small, as current required is less than 1 mA.

Check out the power supply modulator using a basic scope if you have one. Look at the DC voltage output and set the LM317 up for whatever voltage is required—in the case of our varactor-controlled 10 GHz system, this will be +10 volts nonvariable. The varactor is driven with a positive voltage and is DC-adjustable from zero to +12 volts. Verify voltage operation and then use the scope to verify modulation on the AC-coupled scope. A few millivolts is all that is required of AC modulation superimposed on the varactor DC voltage for proper FM modulation.

In operation with the completed system, check all your power supply connections and voltage requirements twice before you connect up the wrong polarity or wrong voltage to the precious Gunn diode and its associated detector diode. The diode can be bypassed with both a small- and large-value capacitor to lower frequency oscillations. You will find that 0.001

and 10 μ F capacitors will do just fine.

The detector diode needs a DC return to draw a little current to bias it slightly on. Most any value small RF choke near 30 μ H or so will suffice. Run shielded leads to both the Gunn and detector diodes. I used miniature coax (RG-174) that was

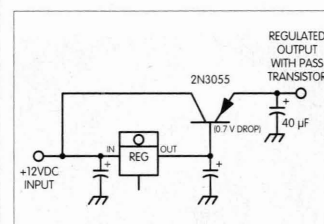


Fig. 4. Bootstrap 2N3055 or similar NPN pass transistor is added to voltage regulator circuit to increase current handling capabilities of voltage regulator. Note: There will be a 0.7 volt drop in regulated voltage out of the regulator due to voltage drop in the base emitter of the pass transistor. Set voltage regulator slightly higher to overcome voltage drop in use.

about one-eighth of an inch in diameter. Coax type is not critical; it's just required for shielding to prevent stray pickup.

Again, I stress: Use different connectors for the connections to feed voltage and detection, to prevent making connections to the wrong lead. If you, for instance, put the detector diode into the 10-volt source, it will destroy the costly detector diode. Use different connectors and you can't make an error in connections.

The detector diode connection is made directly to the 30 MHz input of the Ramsey FR-10 receiver. With the modifications described last month, the receiver should tune over a 400 kHz range of frequencies, making 30 MHz exactly the center of tuning. Normally, you will not have to make any receiver adjustment in frequency. For other stations that might be slightly off-frequency from 30 MHz, you may need to adjust slightly for received clarity.

Operation on microwave is full duplex, just like talking on a telephone. With simple horn antennas, you can communicate over many miles, depending on terrain and path conditions. By adding a small (12 inches in diameter) dish antenna, you can increase available gain by 28 dB

(vs. a small horn, whose gain is about 12 dB). Quite an increase in gain with such a small dish antenna. The same comparison is true for 24 GHz operation. However, a one-foot dish at 24 GHz would have about 35 dB of gain because of its smaller wavelength. As frequency increases, wavelength becomes smaller, and you get more gain for the same area than at lower frequencies. Of course, that's for a dish antenna optimized at frequency.

Well, there you go. The package of the Ramsey FR-10 receiver and the transmitter modulator power supply control circuits should get you on the air with simple wideband FM operation. I tested my circuits using the Ramsey receiver, which proved quite sensitive and of great quality. The frequency I used was 24 GHz, because I knew from past experience that if it worked here it will perform on 10 GHz just as well.

Why pick 24 GHz for a test?

We wanted to complete project testing in time for participation in the ARRL 10 GHz and up contest. I used my 10 GHz narrowband station at home and made several contacts, but I really wanted to try

24 GHz for pure fun and to see if both Kerry N6IZW and I could get operational. I constructed and modified the receivers and obtained some small medical receivers to use for a shielded housing after removing all junk from the cabinet except the fuse and on/off switch. A simple conversion of the cases sure beat the prices of new metal cabinets (hams are frugal at times).

Kerry N6IZW constructed the modulator power supplies, and one evening two days prior to the contest we sat down, bench-tested both units, and got them operational. Kerry fashioned his 10 GHz dish with a small C-clamp, to fix the 24 GHz diode assembly near focus, and that allowed him to obtain quite a bit of gain in his system, possibly as much as 45 dB. I did not have time to haul out the dish feed due to commitment to our grandson's soccer game that Saturday morning, so I used a simple miniature horn antenna less than an inch in area for my antenna. Still, I made contact with Kerry over a short test range of about two to three miles, from Mt. Helix to Kerry's front yard.

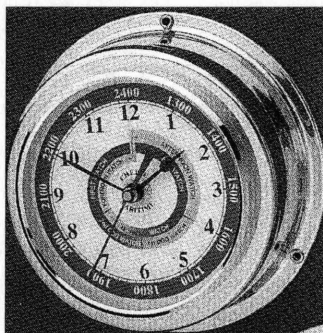
Then Kerry and I met on top of Mt. Helix and communicated with Ed W6OYJ again on 24 GHz wideband FM. He was on top of San Miguel, where there are several television stations and FM radio stations, besides commercial FM repeaters—all co-located near his operation point. We made contact on 24 GHz, but signals were so strong from an interference point that Ed's S-meter was pinned, with or without 24 GHz signals. Both

Kerry and I were able to hear sync buzz from the very powerful video UHF transmitter, even at some 12 miles distance. All in all, it made for a very interesting day and lots of enjoyment.

In retrospect, I can't give enough praise to the Ramsey FR-10 receiver. It delivered in many areas, including the most important one, cost. It is very inexpensive at \$35, and outperforms similar systems. It comes with all component parts, a quality PC board, and easy assembly instructions. In field tests that we ran, it proved to be a very important player, and worked far better than I had hoped. If you haven't picked one up yet, do so if you intend at all to get on wideband FM. You should not pass up this fine bargain.

Next time, I want to get into the test equipment that was constructed to allow our testing at 24 GHz. I will bet your work bench is in the same boat mine was, with nothing above 18 GHz in the testing arena. Well, my old 8551 20-year-old (or older) spectrum analyzer goes to blue light with external mixers, but in reality, it's not very good with regard to what it sees. Next time we'll describe what circuitry was assembled to do quality testing at 24 GHz. The approach is not limited to only this frequency but can be applied to others as well—even lower ones—depending on your testing needs.

The main ingredient needed is a spectrum analyzer that can cover up to a GHz or so. We'll let you in on the plot next month and describe what we came up with. 73 for now, Chuck WB6IGP.



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HOMING IN

Radio Direction Finding

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Build the NorthScope

Sometimes hidden transmitter hunting (T-hunting) is easy. I get sharp, unambiguous bearings, drive directly to the hidden T on good streets through light traffic, and my passengers ask, "What's so hard about this?"

Or was I just daydreaming? Actually, it seems as if every time I take ride-alongs to demonstrate the fun of radio direction finding (RDF) contests, the hunt is extra difficult and I don't do as well as I would like. Of course, it wouldn't be fair to blame the passengers for upsetting my concentration. The fact is that there are wide variations in difficulty of T-hunts due to variations in signal level, polarization, terrain, and the surroundings of both the hidden T and the hunters. I'm always seeking out new ways to get the best bearings in the worst circumstances.

Most two-meter T-hunters in southern California use a yagi or quad of three to six elements on a mast extending vertically from the vehicle window or roof hole. They depend on the radio's S-meter to tell the direction of

strongest signal as they rotate the mast by hand. An RF attenuator keeps the meter on scale when closing in. In urban areas where signals are constantly reflected from hills and buildings (multipath), the beam's sharp forward lobe can pick out each direct and reflected signal component. That's a major advantage of the beam method over a Doppler or dual-switched-dipole setup.

The down side is that interpreting your beam's indications can be tricky and time consuming, slowing you down when speed is of the essence. Sometimes multipath, airplane flutter, and path blockage make the S-meter reading fluctuate constantly as you roll along. Getting an accurate bearing on the direct signal while ruling out reflected signal peaks under these circumstances is not an exact science. A few experienced T-hunters have improved their performance with polar-plot bearing displays of signal strength versus direction on a cathode-ray tube (CRT) or computer.

CRT bearing readouts featured in past "Homing In" columns include those of KK6CU (October and November 1992), NØMKJ (March 1994), and AB6OS/KA6SOX (November 1993). The screen shows signal strength versus direction from multiple sweeps of the antenna so the operator can "eyeball average" fluctuations resulting from motion. It is much easier to discern the most likely direct signal direction in a multipath environment with a polar display than it is with just an S-meter.

North-up is better

All of the above polar displays show signal directions relative to the vehicle's heading. That's fine when you're driving in a straight path and you know exactly which way you're going. But what about T-hunts that take you on winding roads in new housing developments or along desert washes in

the middle of nowhere? How can you interpret the display accurately when it rotates as the vehicle turns?

Last month's "Homing In" showed how a fluxgate compass sensor on your mobile beam detects the mast's orientation with respect to true north. Why not combine the fluxgate sensor and CRT readout to produce a polar display of signal strength that is always relative to north, no matter which way you turn? In aviation terms, this is called a "north-up" display. I call mine the NorthScope.

For good "eyeball averaging," the CRT must display several rotations or sweeps of the RDF antenna at a time. An ordinary oscilloscope won't do—you need a storage-type oscilloscope or a high-persistence CRT like those in radar sets of the pre-computer era. The NorthScope trace of **Photo A** is typical for a strong signal with a small amount of multipath. The higher the beam's gain, the sharper and narrower the major lobe will be. Note that this pattern is consistent for every rotation of the antenna because the vehicle and source are stationary and there are no moving objects such as aircraft in the path.

The NorthScope is at its best in a difficult RF environment such as in **Photo B**. The large repeatable lobe identifies the most likely direct bearing to the T. Reflections and noise in other directions show up as a jumble of non-correlated traces after several rotations of the mast. If there are two or more keyed-down transmitters in different directions on the frequency simultaneously, the NorthScope can resolve bearings for each of them. Try that with a Doppler!

Sometimes the signal level fluctuates, making it hard to find a peak on the S-meter. Worse yet, imagine the S-meter bounce if the signal switches on and off every second or so. (It's legal on some hunts!) Under these conditions, turn the antenna slowly several times, find the peak on the polar display, then read the direction.

A CRT readout need not be a budget-breaker. Used storage scopes show up regularly at electronics surplus sales and swap

meets. They aren't cheap, but if you're an experimenter, you will find other uses for one around the shack. Medical monitors such as the Tektronix Model 603 include waveform storage and are available on the surplus market at lower prices. Look for scopes and monitors that accept two analog inputs and have an X-Y mode, plus gain and position controls for each channel.

Laboratory scopes and medical monitors are designed to operate from household outlets. You will need a DC-to-AC step-up converter to adapt them to mobile use. PowerVerters™ by TrippLite are suitable and available in several wattage ratings. Be advised that the current drain of a storage scope or monitor can be substantial; a heavy-duty car battery is a good idea.

A simple interface

To connect the scope to the sin/cos fluxgate outputs (described last month) and draw a polar plot as the mast goes around, we need a circuit that varies the sine and cosine amplitudes in proportion to the incoming signal level. The analog multiplier, a little-known function block, is perfect for this task. Not to be confused with the more common analog multiplexer (which switches signals), the analog multiplier produces an output signal that is the exact product of two input signals. For instance, if one input is +2 volts and the other is +3 volts, the multiplier output is +6 volts.

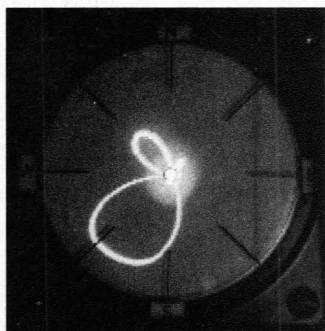


Photo A. A local repeater produces this pattern on the mobile quad and NorthScope. The main forward lobe at 220 degrees true is the direct signal. The second lobe at 320 degrees is reflected signal from nearby foothills.

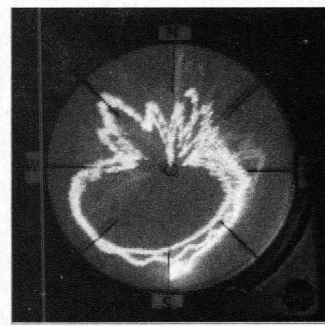


Photo B. This distant simplex signal is on the other side of a hill, so it has lots of multipath and airplane flutter. By viewing several overlaid sweeps with the NorthScope, it is clear that the most likely bearing is 235 degrees.

A four-quadrant multiplier takes into account the polarity of input signals. If A input is -2 volts and B input is +2 volts, output is -4 volts. Put in -1.5 volts and -2.0 volts to get +3.0 volts, and so forth. Typically, the product voltage is internally divided by a factor of 10. This makes sense when you consider that the result of 5 volts times 5 volts is too great for a circuit powered by a 15-volt source if the output is not scaled down.

Analog multiplier integrated circuits are inexpensive and ideal for modulators, wattmeters, voltage-controlled amplifiers, and automatic gain control stages. The major suppliers of these ICs are Analog Devices, Harris Semiconductors, and Motorola Semiconductors. Sin/cos signals swing positive and negative, while S-meter signals are positive. Therefore, the multiplier must operate in two of the four quadrants.

I chose the Motorola MC1495, available from nationwide distributors such as Newark Electronics. The dual-inline package (P suffix) is easiest for home builders to use, but it may be harder to find than the surface-mount (D suffix) package.

Fig. 1 is the fluxgate-to-monitor interface circuit schematic. There are separate analog multiplier ICs and operational amplifiers for the X and Y axes. All parts except the multipliers are available at local parts stores. They fit onto a few square inches of perforated board space (**Photo C**). Construction is simple, even without an etched circuit board. No ground plane is needed, but be sure to put the supply bypass capacitors close to the ICs. Mount S1 and R1 in convenient locations for adjustment during T-hunts.

Inputs of the multiplier ICs are differential amplifier stages. That's ideal for this application because the sin/cos outputs from the fluxgate compass vary positive and negative with respect to a +4.2-volt analog reference. By connecting multiplier inverting input pins to the analog reference, that relationship is preserved through the interface circuit and there is no need for a regulated negative voltage supply. U3A and U3B convert the differential

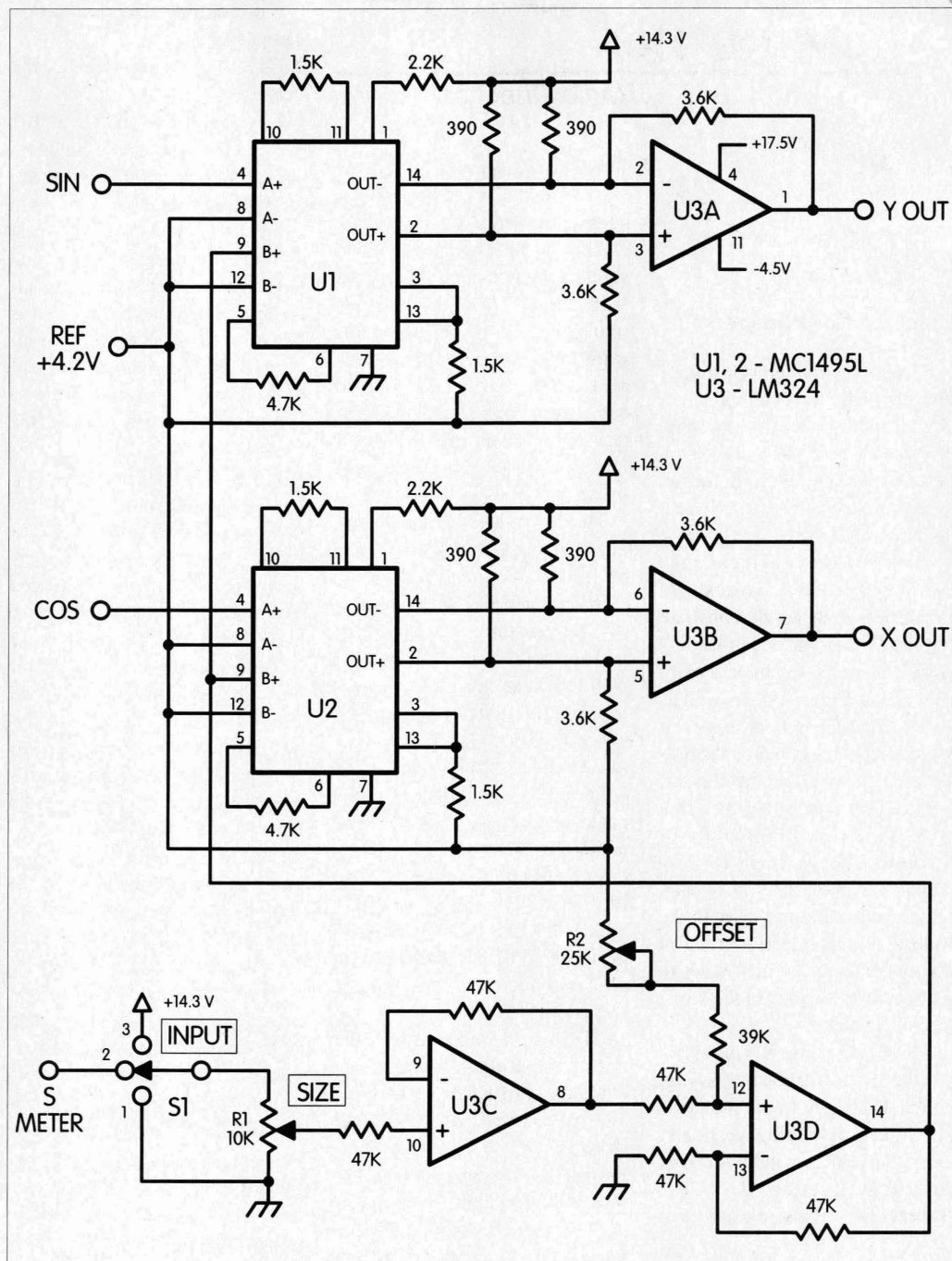


Fig. 1. Schematic of the interface circuit that combines fluxgate compass and S-meter signals to drive a north-up CRT display. Resistors are in ohms and capacitances are in microfarads unless otherwise noted.

outputs of the multiplier ICs to single-ended.

My receiver's S-meter tapoff circuit is described in Chapter 5 of the book *TRANSMITTER HUNTING—Radio Direction Finding Simplified* (available from 73's Radio Bookshop). It includes an operational amplifier to drive an analog panel meter atop the dashboard. The op-amp output in that circuit goes from zero to +10 volts full scale and connects to the point marked S-METER in **Fig. 1**. U3A is a

voltage follower stage with unity gain; U3B adds 4.2 volts to the S-meter voltage as required by the differential analog multiplier inputs.

To ensure sufficient headroom in the multiplier and avoid waveform clipping under all input conditions, use care in choosing supply voltages for the interface circuit. The positive supply to the MC1495 ICs should be regulated. Supplies for the LM324 quad op-amp can be unregulated. The LM324 positive supply (pin 4)

range is +15.8 to +20 volts; negative supply (pin 11) can be -1.5 to -12 volts. Look for places to tap off these supply voltages within your monitor.

Tune-up is easy

A few simple alignment steps are required to set up the NorthScope for the first time. Connect all the assemblies and install the fluxgate sensor on your antenna mast. Set the focus control on your monitor for a minimum-size spot. Keep the

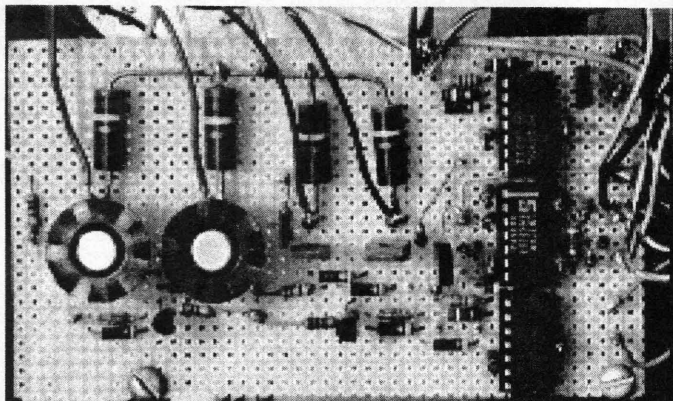


Photo C. The three ICs and associated resistors on the right side of this perforated board are the fluxgate/S-meter interface components. The rest of the board holds CRT deflection amplifiers for the home-built monitor.

brightness control as low as practical to avoid burning the phosphor screen of the scope. X and Y channel sensitivity settings should be about one volt per division, depending on your screen size.

Turn SIZE control R1 on the interface to zero. Set input switch S1 to SPOT (position 1). Set the X and Y position controls on your scope to center the spot on the screen. Turn the mast and sensor around 360 degrees; the spot should not move. If it does, adjust OFFSET control R2 until the spot does not move as the mast turns. Now re-center the spot, if necessary, with the position controls.

Set S1 to CIRCLE (position 3). Adjust the SIZE control to move the spot about an inch toward the edge of the screen. Rotate the mast/sensor and adjust the X and Y gain controls and the two screwdriver-adjusted potentiometers on the left side of the fluxgate compass display for a perfect circular pattern around the center spot position. The E-W and N-S pots on the fluxgate equalize the left-right and top-bottom voltage swings such that positive excursions equal negative excursions for each axis. This makes the circle trace exactly over the spot (see Fig. 2). If the circle is elliptical instead of round, adjust the X and Y channel gains as appropriate.

Set S1 to TRACE (position 2). With your receiver tuned to an active repeater and enough RF attenuation that the S-meter does not peg, swing the beam around and adjust the SIZE control for a

trace similar to **Photo A**. There is plenty of gain in the S-meter channel, so you can get a full-size trace even if the incoming signal only moves the S-meter to quarter-scale.

Notice that the reflection lobe in **Photo A** is quite sharp but the major lobe is broad at the outside. This is caused by nonlinearity in my receiver's S-meter circuits. Normally this is not a problem, but I can get a sharper lobe if needed by adding RF attenuation to the receiver and turning up the SIZE control to compensate. It should also be possible to add nonlinear devices such as diodes to the feedback path of U1A to make that stage an antilog amplifier; I have not tried this yet.

While the north-up display could be used with a motorized mast like the displays of KK6CU and others, I prefer turning the mast by hand. Not only does it simplify the project (no slip rings for the coax and sin/cos signals), it also makes it easier to find the exact direction of signal peaks as displayed on the scope. I can rock the beam back and forth until I'm satisfied that I have found the optimum bearing. This is particularly helpful when there is airplane flutter or rapid amplitude changes due to keying or high modulation levels.

I have enjoyed building several oscilloscopes over the years so I decided to make my own monitor/display for the NorthScope. By modifying a commonly available "boat anchor" scope and raiding the junk box, the cost was

under \$100. The monitor draws about as much as a typical mobile receiver, so I don't worry about draining the car battery. Next month's "Homing In" will have all the monitor details and

some thoughts on computerizing the NorthScope. Meanwhile, keep sending your T-hunting news and column suggestions via E-mail or my post office box. Thank you! 73

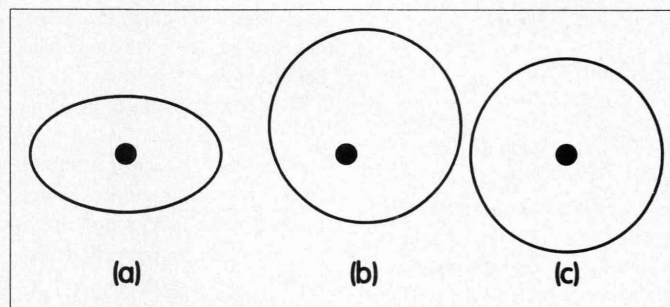


Fig. 2. At (a), unequal horizontal and vertical gain. At (b), offset due to misadjusted E-W and N-S controls. At (c), a perfect circle in proper position.

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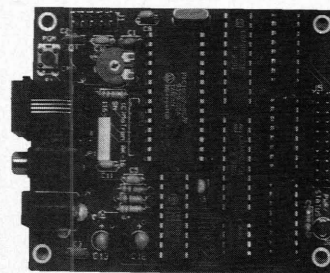
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CIRCLE 248 ON READER SERVICE CARD

QRX...

continued from page 8

Shuch. He has just returned from a two-week lecture tour of Australia and New Zealand, as part of his three-year-old effort to inform and involve radio amateurs around the world.

SETI scientists seek to determine through microwave measurements whether humankind is alone in the universe. Since Congress terminated NASA's SETI funding in 1993, The SETI League and other scientific groups have been attempting to privatize the research. Experimenters interested in participating in the search for intelligent alien life, or citizens wishing to help support it, should visit us on the Web at [http://www.setileague.org/], E-mail to [join@setileague.org], send a FAX to (201) 641-1771, or contact The SETI League, Inc. membership hotline at (800) TAU-SETI. Be sure to provide us with a postal address to which we will mail further information. The SETI League, Inc. is a membership-supported, nonprofit [501(c)(3)], educational and scientific corporation dedicated to the electromagnetic Search for Extra-Terrestrial Intelligence.

From a press release from The SETI League, Inc.

Reach for the Sky!

Hams in Mason County, Washington, won't be encumbered by a telecommunications ordinance that would have restricted the height of ham radio towers to 70 feet and imposed other regulations. Andrew Forsberg WV7M, of Grapeview, reports that several hams in the largely rural western Washington county (population approximately 30,000) cited federal preemption over local regulation of amateur radio activities as well as the hobby's public service dimension to get the County Board of Commissioners to exclude ham radio from the new law last month.

Forsberg says that, as originally drafted, not only would tower heights have been limited to 70 feet, but building-mounted towers could not have been more than 20 feet tall. It also would have held the height of vertical antennas (called "whips" in the proposed regulations) to 15 feet. In addition, the proposed law would have required landscaping to hide a tower and lot setbacks equal to a tower's height—something often impossible on a small residential lot.

Forsberg said that, at first, the drafters of the new ordinance "seemed to be unmoved by amateur radio considerations." In addition to federal preemption, the county hams pointed out the connection between a good antenna system and ham radio's public service contributions—in an area subject to floods, earthquakes, power outages during winter storms, and even volcanic activity. Commissioners "began to soften their position," Forsberg said.

"By the time the vote was taken, the commissioners were well-informed of our position and were commending amateur radio for its outstanding contributions to the community," he added.

From the *ARRL Letter*, Vol. 17, No. 6, by way of *PARKing Ticket*, newsletter of the Plano (TX) Amateur Radio Club, March 1998.

10 Sure Ways to Destroy Your Club

1. Stay away from meetings. Show up at meetings only if you have problems or complaints.
2. Decline office or appointment to any committee. Then get angry if you are not nominated.
3. Insist on describing the club in negative terms only.
4. Never prepare an agenda, never plan the details of an event, and never coordinate with affiliates.
5. Don't do any club work if you can avoid it. Then, when the old reliables pitch in and get the job done, accuse them of being a clique.
6. If you do come to a meeting, don't speak until the meeting is finished. Then criticize and say how things should have been handled.
7. Oppose all new programs as being a waste of the membership's money.
8. When nothing new happens, complain that the officers lack imagination.
9. Read your bulletins infrequently. Then complain that you don't know what's going on.
10. Never introduce yourself to new members or visitors—make them come to you.

By John E. Matheson, *Board Briefs*. We borrowed it from December 1997's *Modulator*, newsletter of the Fort Myers (FL) ARC, Earl Spencer K4FQU, editor.

Russian Device Could Make Bombs Impotent, *Jane's Defence Weekly* Reports

Washington DC, March 19, 1998. Russian scientists have developed a compact high-current electron accelerator that could potentially stop car engines and destroy the electronics arming and firing circuits of bombs, *Jane's Defence Weekly* reports in its March 18 issue.

The device, called RADAN, is smaller than an attache case, weighs about eight kilograms, has a directional antenna and a 12-volt rechargeable battery. Dr. Ira Merritt, of the advanced technology directorate at the US Army Missile Defense and Space Technology Center, told the Congressional Joint Economic Committee recently.

Merritt said the device can stimulate the output of lasers, x-rays, wideband radio frequency and high-power microwaves. RADAN has a pulsewidth of two nanoseconds and an output pulse bandwidth from 1 MHz to 5 GHz.

JDW reports that US Army scientists have long been monitoring Russian, French and Swedish programs in this area to more fully understand the vulnerability of US battlefield electronics and computer systems to RF weapons.

"There is an increasing variety of equipment capable of generating very short RF pulses that are capable of disrupting sophisticated electronics," Merritt said.

These types of pulses, which may last a nanosecond, are not considered in many current electronics design standards and will require new forms of RF and electromagnetic interference protection.

Jane's Information Group, which publishes *Jane's Defence Weekly*, was described by the CBS program *60 Minutes* as "the closest thing there is to a commercial intelligence agency." Jane's is the leading provider of defense, aerospace, aviation, transportation, geopolitical, and police and security information to the world's militaries, governments, universities and businesses.

From a press release by Jane's Information Group.

Paraguay Orders Radio Station Off the Air

The government of Paraguay has ordered a radio station off the air for airing a phony news program that simulated a coup in progress in the country.

Radio Uno, in the capital city of Asunción, was ordered closed by the state's National Commission of Telecommunications, under a law which establishes that broadcasting must be based on the transmission of objective news from responsible sources. The program faking a coup came just two days after rumors of a real coup circulated throughout the Paraguayan capital.

From shortwave reports, via Bill Pasternak WA6ITF at *Newsline*. 73

The Fun Radio

continued from page 20

speaker or stereo headphones to an adequate audio level. As an option, of course, the Fun Radio could be connected to the phono input of a hi-fi audio amplifier system.

Conclusion

In years past, a longwire and ground provided the necessary signal input for the radio. With the options available today, most anything will work, but a longwire should be used for the fun of it. Also, be sure to use a good earth ground.

Build the Fun Radio and enjoy a step into the past with the fun of today. Put on the headphones—twist the regeneration and tuning dials and listen to the squeals and sounds of yesteryear! 73

VHF and Above Operation

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Bits and pieces for microwave and VHF

This month I would like to address some frequently asked questions concerning microwave hardware, such as: What do I need to pick up to start a microwave test bench? Which components of the test bench are most used for basic work? Also, I want to cover some simple test devices that can be constructed for both microwave and VHF operation.

A basic work bench for microwave should have some means of measuring power capable of displaying low levels of power. Here on microwave we talk of powers of -10 dBm and +15 dBm power. This is in difference of power on a VHF or UHF band where power can be measured in watts of power. Power in microwave is quite dear, in that higher levels of power are precious and expensive to obtain.

There is a direct comparison between power expressed in watts and dBm. It's not hard to remember a few simple facts to keep in mind when comparing the two power levels. For instance, two watts of power can be measured on a microwave power meter to a very fine accuracy, as long as appropriate attenuators are used. Most power meters for microwave are rated to +10 dBm maximum, so to measure two watts of power, which is the equivalent of +33 dBm of power, an attenuator is needed to reduce the power applied to the microwave power meter. It's just simpler to use dB vs. watts of power when making calculations. A simple rule to remember is: For every 3 dB of increase of power, you have

doubled your original power in watts, or a portion of a watt.

As I stated, most microwave power meters are not capable of measuring power over +10 dBm maximum power or the sensitive power head will be destroyed. To allow the power head to work at much higher power levels, you insert precision attenuators in between the power source and the power head, reducing power to an acceptable level below the +10 dBm maximum. For instance, to measure a two-watt power source the level in dB would be +33 dBm. Inserting a 30 dB attenuator would reduce the power to 3 dBm, as read on the power meter, with the 30 dB being reduced in the attenuator. To obtain the true power measured, add the attenuator and the reading on the power meter (+3 dB) to obtain the true reading of +33 dBm.

Let's back up a moment. Remember, we said that +33 dBm was two watts, and as its power doubles, it increases 3 dB for each doubling of power. Therefore, 1 W = +30 dBm; then 2 W = +33; 4 W = +36; and 8 W = +39 dBm; 16 W = +42 dBm. Another good rule to remember is that 1 W = +30 dBm and 10 W = +40 dBm and 100 W = +50 dBm and so on.

To allow you to measure these power levels, you need a good attenuator capable of attenuating power and handling power dissipated in the attenuator, to make a power meter reading. If you wanted to measure a 10 W microwave transmitter or a similar VHF transmitter the method is the same: 10 W = +40 dBm, and the power meter is capable of +10 maximum. Therefore you must attenuate at

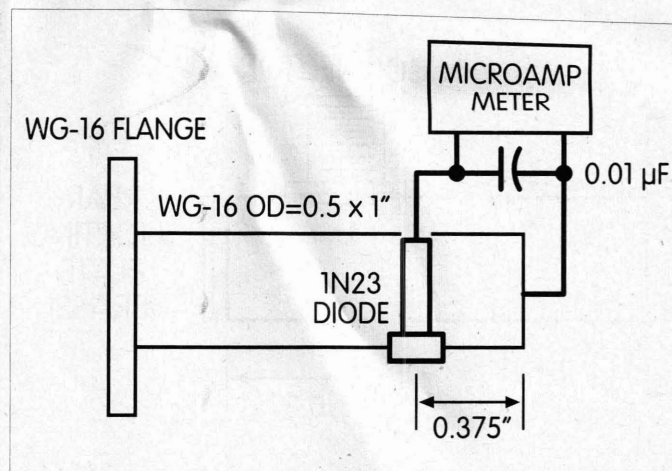


Fig. 1. Microwave Detector mount for 10 GHz showing detector diode spaced 1/4 guide wavelength from rear wall of detector mount. Can be home constructed from scrap waveguide and waveguide flange. Insulate top of 1N23 detector diode. Mount suitable meter very close to diode and bypass with .01 capacitor. For increased sensitivity, insert meter amplifier between diode and meter.

least 30 dB, or, better still, 40 dB of power before hooking up to the power meter. Making a reading in this, you would combine the reading on the power meter (either negative or positive) in reference to zero dB on the meter to calculate final power observed.

Measuring 10 watts with a 40 dB attenuator, the power meter should read zero dBm, or 10

watts attenuated to zero dB by a 40 dB attenuator. Similarly, if you had used a 30 dB attenuator you would expect to read +10 on the power meter for the same 10 watts of power. While this will work, it is a bit risky, as it is the maximum power you should be absorbing in the power meter detector head.

A good set of attenuators to have on hand for making power

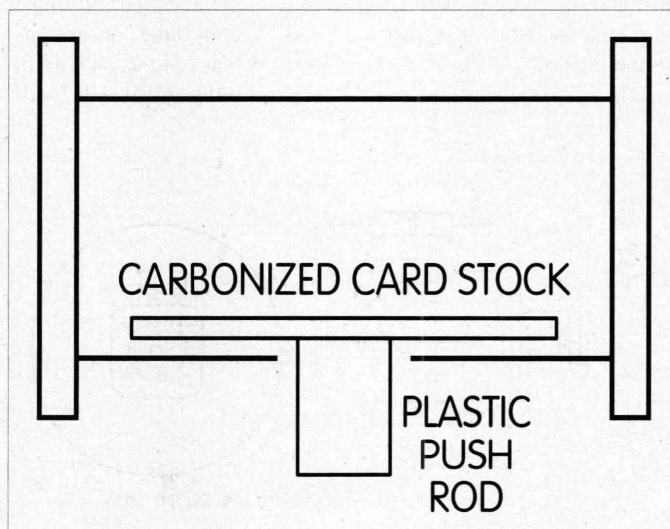


Fig. 2. Waveguide attenuator construction typical of both the fixed and variable types. The fixed can be useful but the variable waveguide types are much more desirable, especially small size variable attenuators. The waveguide attenuator is open on either end; as card is inserted toward center of guide, attenuation is increased. Attenuation is near zero when resting near bottom wall of WG.

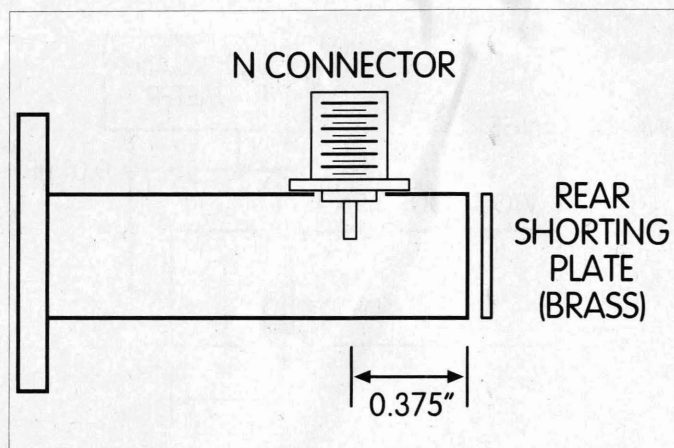


Fig. 3. Waveguide transition. Very similar to detector mount but uses probe from coax connector penetrating waveguide center to pick up microwave energy and convert it from waveguide to coaxial. Typically type "N" connectors are common. The connector is mounted on top of the side dimension of waveguide; drill hole to clear N coax connector flush with bottom of connector and waveguide. Space .375 inch from back of rear shorting brass plate. Solder all parts together.

measurements includes 3 dB, 6 dB, 10 dB, 20 dB and 30 dB. Normal microwave attenuators look much like a coaxial connector with a male connector on one end and a female connector on the other end. This is so that several can be put in series with each other and added together for total attenuation. For the basic types of attenuators two watts is normal. For increased power levels attenuators will have an additional heat sink or rings for power dissipation around the body of the attenuator. These types of attenuators

are much larger and typical power dissipated can be 60 watts to 100 watts.

Another factor in attenuators is their internal structure that limits frequency response. Not all attenuators are created equal. Some attenuators in the two-watt range are good from DC to 12 or 18 GHz. Others are constructed in such a way they function well in the 2 to 4 GHz area but become unpredictable above these frequencies. The best idea is to read the label on any attenuator you want to pick up in surplus, and see if it will func-

tion at the frequency range you intend to use it at. The golden rule is "If it's expensive and unmarked, check it out very carefully before purchasing." Don't buy something you can't use.

One way to test to see if an attenuator is alive is to carry an inexpensive VOM in your pocket. A lot of attenuators sold are destroyed by applying over-power to them, cooking the resistors inside. Check to see if you have a DC resistance from input to output on the center pin. Readings should be in the low-to-mid 100 or so ohms. Then check from each end's center pin to case ground. The readings on each end should be similar, and again, in the low-to-mid 100 ohm ranges. For example, a typical "T" type attenuator would measure 50 ohms from input to output and 61 ohms from each end to ground.

Waveguide devices

The most useful waveguide devices that an amateur can use are threefold: a detector mount, a transition, and a variable waveguide attenuator. These devices need be considered only if you intend to work with waveguide. Some devices that can use waveguide fixtures are traveling wave amplifier tubes (high power at microwave), Gunn diode units for wideband FM on 10 GHz and most bench test equipment associated with waveguide fixtures.

The most important item to obtain for waveguide use is the transition that converts a waveguide flange connection to a coaxial connector. This allows interweaving with waveguide in certain pieces of equipment you might have and coaxial connections in the rest of the equipment. Typical useful interconnections are antenna feeds, like waveguide horns, and other antenna hardware.

In this testing arena, a good waveguide detector mount can be put to use on a remote antenna such as a horn or other broadband device. In practice,

horn and detector mount are cabled back to a testing site some 100 to 200 feet in front of the test antenna and detector for indication at the transmit site. Now, you have a simple part of an antenna test range to show dish aiming and a relative gain measurement. It's a good idea to try simple tests and methods to find out how well systems are functioning, using such equipment.

Waveguide attenuators, especially variable attenuators, can prove useful by reducing a signal at the source, to see how far a signal can be reduced before you can no longer copy it, giving some means of finding minimum signal level of detection, another method of comparison to try out using simple equipment.

Waveguide and coaxial attenuators work the same, but in different styles and package forms. The principle is the same in either domain. In a coaxial domain the ends of a connector can be soldered to common small composition resistors, duplicating the values needed for a similar attenuator type. These will function well as an attenuator, but at a very reduced frequency before they become unreliable. The problem with this type of construction is that it's OK for six or two meters, and *might* work at 450 MHz as well, but as the component size starts to become a part of a wavelength at the intended frequency of operation, the performance falls apart very fast.

Components' size and physical construction techniques need to be quite special if they are to function with accuracy at frequencies above several GHz, let alone function to 18 GHz. Normally the resistors used in commercial attenuators suitable for high GHz work are constructed with special-shaped cores and have the resistance material deposited on them. The combination of shape and deposited resistance material give the very special characteristics (low SWR) at microwave frequencies.

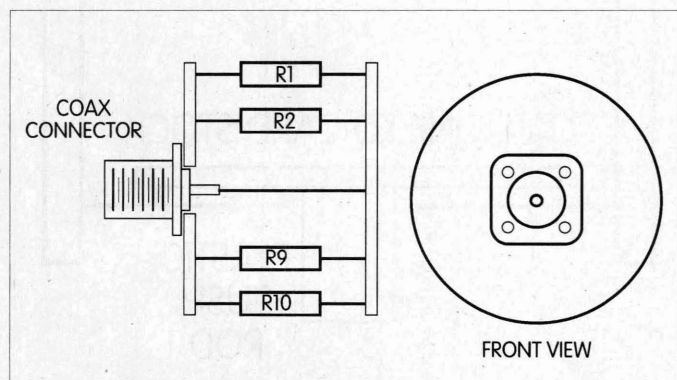


Fig. 4. Dummy load construction showing means of construction for VHF operation from multiple resistors equaling 50 ohms or nearly 50 ohms, placed between two brass or copper plates. The hole in the center of the top disk is for coax connector. Solder to top plate; center conductor of connector is soldered to bottom plate. Unit can be inserted into a small can for shielding after construction.

Attenuators of the waveguide type are constructed with a card inserted in the narrow side of a waveguide and have deposited resistance material on one side of this card material. It can be thought of as being cardboard with graphite on one side of the card for a mental picture. When the card is adjusted to minimum attenuation the card rests against the short side wall of the waveguide. As the adjustment arm is changed in length to more attenuation the card is pushed out from the side wall into the waveguide towards the center of the guide. As the attenuator is pushed or adjusted towards the center of the waveguide the attenuation increases rapidly.

Transmission of RF in a waveguide results in the most RF signal at the center of the waveguide. So anything inserted into the waveguide at this point will intercept a great deal of the applied RF, reducing the output of the system after the attenuator.

50 Ω dummy load

Home construction of attenuators and detector and transition mounts is possible. However, in the case of attenuators, special techniques are needed, allowing good operation in the very high microwave frequency ranges. Home construction is not recommended. I stated before that attenuators and dummy loads can be constructed from common carbon resistors. While it's hard to find an exact 50 Ω resistor, several higher-value resistors can be placed in parallel to obtain a value of near 50 Ω .

Simple 50 Ω dummy loads (terminations) can be easily constructed from a large number of high-value 1 or 2 W carbon resistors. Let's say, for instance, if you want a 50 Ω 10 W load for two meters. Calculate: If you had 10 each one-watt 500 Ω resistors, placing them all in parallel with each other, you would have exactly 50 Ω with a power dissipation of 10 x 1 W ... exactly what you wanted.

In the real world things are not all that neat. But calculate with a standard value like 680 Ω (one watt) by placing 10 in parallel. That would be 68 Ω ; 14 resistors result in 48.5 Ω , and 13 resistors result in 52.3 Ω . See how simple it can be to come up with a 50 Ω dummy load for test?

To find out if you have suitable resistors to total the correct value for a 50 Ω load, just divide the value of resistors by the quantity of resistors. An easier method is to take a value you have on hand and divide it by the value you wish to arrive at, and the result is the number of resistors in parallel needed. (Example: 750 Ω stock resistors divided by 50 Ω desired = 15, the quantity needed in parallel to obtain a 50 Ω final product.)

Waveguide transition

Now what about the transition mount for waveguide that I mentioned earlier? All it needs to be is a short section of waveguide, about an inch long, with a coaxial connector mounted on top of the wide face of the waveguide. Usually a type "N" connector is used to be the coaxial interface to the waveguide. It is mounted on the wide side of the waveguide centered on the broad face of this side. The center connector (solder terminal) is extended into the hole in the waveguide and into the inside of the guide. This pin (center of coax connector) is spaced 1/4 guide wavelength from the closed end of the rear wall of the waveguide. This dimension is approximately .375 inches for 10 GHz operation.

A guide wavelength is similar to figuring a wavelength in coaxial cable, when its length is corrected by the velocity factor of coaxial cable. Just for the record, a guide wavelength (X_g) is equal to 1.5 inches for 10250 MHz making 1/4 X_g equal to .375 inches. Now, if you use an "N" connector it can be placed on the waveguide wide dimension with its flange just extended over the rear of the guide equal to the thickness of

QRP

Low Power Operation

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Ham radio takes to the fields and back woods this month as Field Day once more brings the hams out of the woodwork. If you've never operated Field Day running QRP, you're in for quite a thrill!

A QRP Field Day

Perhaps one of the best things going for a QRP Field Day is the ability to carry everything you need in a backpack! Here's a list of the things I take with me for QRPing in the field:

- A QRP rig. I like a multi-band rig that covers at least the 40 and 20 meter bands. If possible, 80 meters comes in real handy for late-night contacts.

- An 18 amp/hr gelled lead-acid battery—fully charged, of course. One of my expanded analog voltmeters to monitor the

battery's voltage goes in the Field Day kit, too.

- To keep the battery topped up before the 2 p.m. start time, a 10-watt Solarex Lite™ solar panel and a Micro "M" charge controller finish out the power system.

- There are only two antenna choices I will ever use for Field Day. One is the G5RV multi-band dipole; the other, a random-length wire. To use either one, an antenna tuner will be needed. The antenna should be ready to go on site. You don't want to waste time assembling an antenna when you could be drinking Diet Coke® and munching on Oreo® cookies! Of course, you'll need some support rope and perhaps an extra antenna insulator or two, just in case.

Continued on page 74

the flange. When the shorting plate is attached to the rear wall under the connector's flange it makes for a flush mounting.

Both the detector mount and the transition are constructed quite similar in that the active element is spaced 1/4 guide wavelength from the end of the rear wall of the waveguide. Shown in **Fig. 1** is a simple detector mount that has the cartridge-type diode (1N23-type) inserted from the bottom of the waveguide (grounded to WG at this point). The top of the diode passes through the top of the WG through a small hole, and is built up with insulating material just below and on top of the diode, like cellophane tape washers over the diode's pin. The pin is insulated from touching the

sides of the hole and connection is made to the pin with as short a connection of coax cable to the meter as possible, which is bypassed for RF with a .01 capacitor. An amplifier can be used for improved sensitivity by connecting it between the meter and the diode. Normally this type of detector is used on wavemeters and similar-type devices for indicating a relative RF measurement. Construction methods are not critical.

Next month I want to present some items you can put together in your garage that are useful in demonstrating RF transmission on a simple antenna test range. I will get into how SWR is measured at microwave compared to VHF and describe some test and measurement equipment used in these tests. 73 for now, Chuck WB6IGP.

ABOVE & BEYOND

Number 21 on your Feedback card

VHF and Above Operation

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Crystal Oscillators and Frequency Standards

Building on last month's column covering frequency measuring equipment, I want to cover crystal oscillators, especially the frequency-standard type of oscillators. In equipment today, ranging from the very basic VFO-controlled HF transceivers to the more sophisticated transceivers on the market, all are controlled by some form of crystal oscillator, if only a 100 kHz calibrator in the most modest of rigs.

The heart of any system rests in its ability to determine frequency, or at least allow the operator of the system a means of knowing where you are in respect to frequency. After all, if you don't know where you are, how can you make easy contacts with others? Some very simple QRP rigs can operate without a sophisticated receiver dial mechanism, but they do have the provision for setting frequency approximately by "NETTING" the transmitter frequency to the receiver. In these very simple and compact systems, the transmitter is

crystal-controlled and probably shirt-pocket-sized, with very few frills incorporated.

The heart of the entire system is a CW transmitter that is crystal-controlled at the operating frequency. For stable design, the crystal oscillator needs some basics to ensure good clean operation, including voltage regulation of the battery supply voltage, usually 12 volts. It's a good idea to regulate down to a much lower voltage to ensure good headroom for the voltage regulator. This allows the battery to run down but still have the voltage regulator functioning at this lower voltage. The oscillator does not see any voltage change. Usually 7 to 10 volts is chosen for the oscillator-regulated supply voltage.

Other considerations include using a circuit for the crystal oscillator that does not try to run a high crystal current and draw maximum power out of the circuit. High current in the oscillator circuit is counterproductive, and just heats things up, making for unstable circuitry and possibly a fractured crystal. It is better to run crystal oscillators at very low power and current draw, with loose coupling. This allows the following circuits to do the work once the oscillator is doing its job. The oscillator stability im-

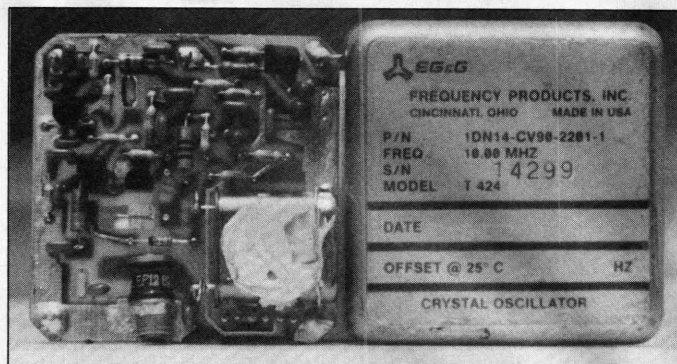


Photo A. TCXO oscillator: 10 MHz, temperature controlled.

proves when other stages do the heavy work, amplification and isolation. Usually these stages are called buffer amplifiers.

Let's take an example from an HF transceiver and see what kind of stability is required for good operation and how it applies to crystal oscillators. In this example I will use a worse-case scenario at maximum frequencies. Let's assume our rig is operating at 30 MHz and the main oscillator is a synthesizer deriving its main clock from an onboard crystal oscillator. This onboard oscillator could be at almost any frequency from 1 MHz to 5 or 10 MHz, or even at an odd frequency such as 3.579545 MHz (TV chroma crystal).

In any case, let's assume the basic stability of the crystal to be 1 Part Per Million (1 PPM). In the 30 MHz transceiver described above, that would

make the system frequency error at 30 MHz about 30 Hz—not bad for an HF rig. Even if the error were in the 300 Hz range it would fall in the SSB passband of operation and would not be noticeable.

I checked several HF rigs that were manufactured in the last 10 years and found a frequency disparity between the digital frequency display dial and the actual frequency, an average error of 400 Hz between the three rigs tested. This certainly proves that for 30 MHz or lower operation, a semi-modern rig is quite accurate for HF operations, and that is without any calibration since the rigs were new some 10 years ago.

Now you can see we have added another stability factor: crystal aging. As time goes on, all crystals need to be reset as even the most ultra-stable crystals and exotic circuitry do indeed drift

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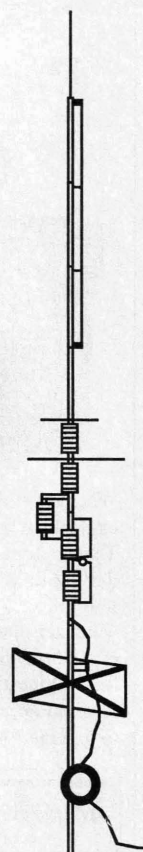
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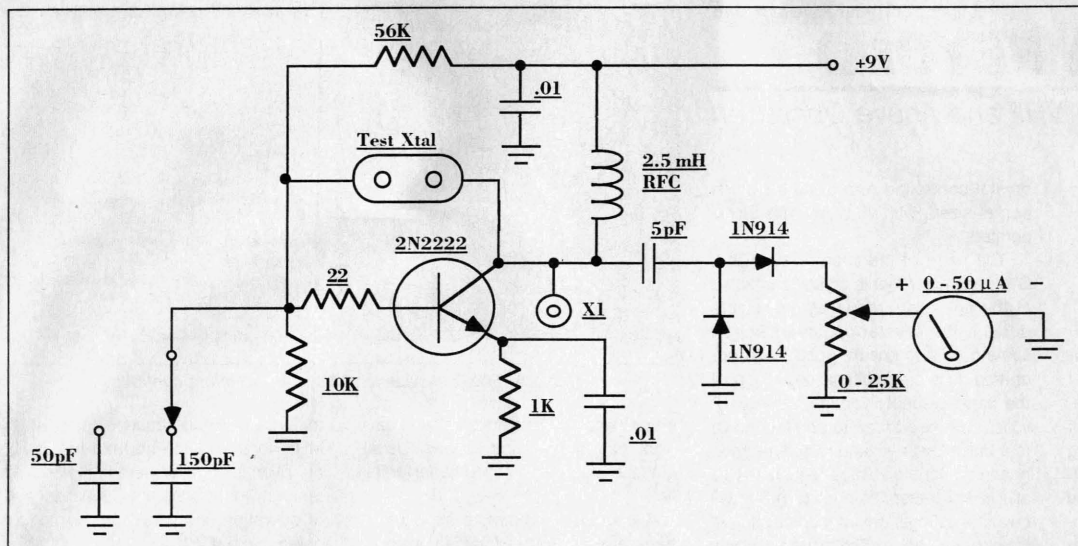


Figure 1. Sample crystal test oscillator. Diodes: 1N4148 or 1N914; RFC: not critical value; X1: connector for 6" antenna wire, providing RF to the receiver for testing/monitoring.

due to aging. This might not seem much at the present level of discussion but it does happen. New crystals seem to age (change frequency) faster than older (burnt-in) crystals. The change is quite small in comparison to the 1 PPM we have been talking about, and in most cases need not concern you at all in the context of the HF scenario above. Basically, look at aging as a point of interest in very high-stability oscillators/

standards. Note that periodic calibration is required in all applications to maintain very accurate conditions.

Let's go back to the HF system error scenario, assuming a 400 Hz offset from indicated frequency (I call that excellent for HF operation). So what's all the excitement about being off frequency? Well, when this kind of error is measured at higher frequencies it can become out-of-hand in short order. If we

transfer the same error, 400 Hz at 30 MHz, to 148 MHz, it becomes almost 2.5 kHz. That's one-half the FM bandwidth for standard operations, and begins to become a gross error. If the same factors are applied to 450 MHz or even 1.2 GHz the error starts to become quite unmanageable. We know that these errors are not present in our modern systems at 450 or even at 1296 MHz. How is that accomplished? Well,

these systems use a tighter frequency standard (clock) for the PLL synthesizer internal to the system and systems that do not use frequency multiplication, which increases error.

These tighter frequency controls are present on almost all of the modern-day synthesized systems for almost any frequency today. Now, how does this compare to home-built oscillators using surplus crystals, or test evaluations of older commercial oscillators purchased on the surplus market? Let's use price as a beginning equalizer and see where that leads us. Aside from individual quartz crystals, either surplus or new, installed into home-constructed oscillators, the output frequency can vary quite a bit. How much depends on the circuitry of the oscillator and how it interfaces with a crystal and its series and parallel parameters and load capacitance.

If the crystal is surplus, not much is known about its particular characteristics (as described above). Construction with these components can be fun but will need tweaking, and possibly a different type of circuit to make the crystal operate on the exact marked frequency. In other words, plugging a crystal into a standard oscillator will make it oscillate but not at its specific frequency. For that

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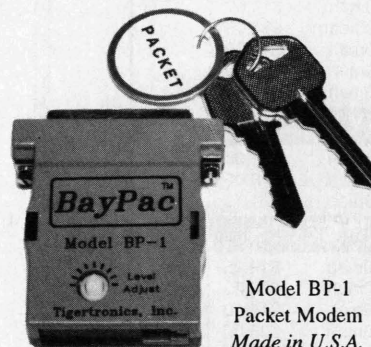
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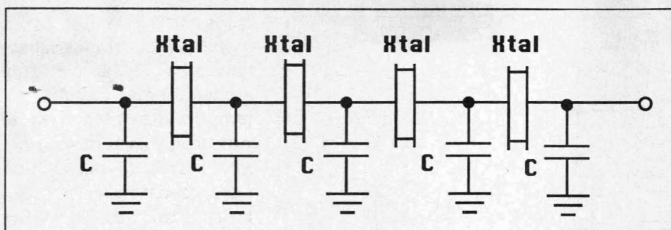


Figure 2. Crystal ladder filter, constructed from four matching crystals. Capacitor values are approximately 15 to 30 pF.

to take place, both the crystal and the circuit have to be in concert with each other. See Figure 1 for a standard-type test oscillator to evaluate a crystal's functioning.

Crystal-matching for a filter application is done with a signal generator and a special piece of equipment called a vector voltmeter. This vector VM notes phase changes in respect to frequency, which is varied for this test. In this way you can match the exact characteristic of the series and parallel resonances of each crystal under test.

A simpler test setup is to use the simple oscillator shown in Figure 1 and listen on a receiver where the crystal oscillates and without readjusting anything but the receiver. Test other crystals of the same marked frequency. When you find a bunch of these, some measure of assurance is a given that these crystals are of nearly identical frequency. Use the receiver frequency for

this match test of the tested crystals. Put the crystals in a row according to receiver frequency. When you have a bunch of crystals of the same characteristics, use them for the filter. The other crystals are good, but not matched. Try this test and you will see the frequencies of the crystals vary quite a bit. See Figure 2 for a typical crystal ladder filter.

Other types of oscillators include "PUT UP," as I call them, or complete oscillators and crystal assemblies. These units comprise a complete oscillator, including the crystal, and are assembled into a package. The simplest and most inexpensive is the TTL IC-chip-socked oscillator. This device has no external calibration, is powered by +5 volts, and can be inserted into a 14-pin IC socket for use. They cost less than \$5 each new and I have found them in surplus, removed from PC boards, for about 25 cents each. Specif-

ic frequency of these is almost endless, especially in surplus. Mouser Electronics lists these oscillators from 1 MHz to 64 MHz, with about 25 stock frequencies. The stated accuracy is ± 50 PPM, or 1 kHz at 10 MHz. That's not too bad, considering that new it costs less than \$5. A 30 MHz oscillator connected into a 10 GHz detector mount makes a "BOOMERANG" for testing a 10 GHz wideband FM system. See Figure 3 for the BOOMERANG construction details.

The next oscillator models that have become available through surplus take a big jump in accuracy. They differ in several ways from the free-running non-adjustable TTL type unit just described. Most of these units are designed around two different circuit concepts for stabilized operation. Both methods incorporate lightly-loaded multi-stage circuits with tight voltage regulation to minimize external effects on oscillator stability. This type of circuit has adjustment capability to trim to an exact frequency by external means. Most are mechanical screwdriver adjustments, but some can have external voltage/varactor type control.

These two types of oscillators differ in how control is applied to the entire unit to maintain a high state of accuracy. The first type uses an oven-type heater to heat the entire crystal and oscillator circuit to a fixed regulated specific temperature. This temperature is far above ambient room temperature and is usually somewhere near 125 to

140 degrees F. The crystal and its associated components are selected for operation at this fixed temperature. This circuit is referred to as an "Oven-Controlled Crystal Oscillator," or OCXO. As such, the oven circuitry requires a lot of current to warm the entire unit, bringing it to proper operational temperature.

Fast warm-up oscillators' current could be 24 volts at 1 amp before the thermostat kicks in and throttles the current back to, say, 1/2 amp or so. An oven circuit can be a power transistor with lots of current to generate heat, coupled to a heat sink (the oscillator compartment), or resistance wire-wound over the oscillator's inner case. Both produce heat to the OCXO circuit. High-quality OCXOs can cost from \$300 to \$1,150 for 0.1 PPM or better oscillators. Prices vary, with costs about 10% of new for surplus oscillators. See Photo B for a sample OCXO.

Where to find high-accuracy OCXOs? Well, almost all surplus military test equipment uses a very high-quality oven-controlled oscillator as the main system clock. Sure, they do draw a lot of power from the supply, but when you are plugged into 110 mains it is just another trade-off in circuit design. The designs vary quite a bit. I have opened up about 20 different models from about half as many manufacturers, and the circuits vary wildly. Some of the very expensive types utilize several ovens for premium temperature stability. The idea here is to have a very slow warm-up

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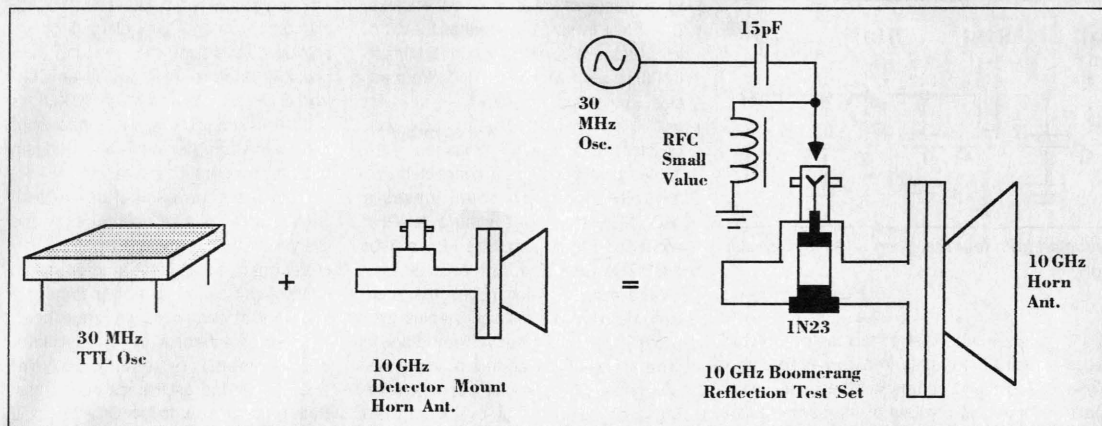


Figure 3. 10 GHz BOOMERANG using a 30 MHz oscillator. A 10 GHz BOOMERANG test set for use with wideband 30 MHz IF systems, providing enough signal to a transmitting 10 GHz system.

cycling oven controlling the internal oscillator circuit. This entire circuit oven and oscillator is then insulated from temperature changes with a styrofoam package and this is contained within another oven-controlled circuit. The result is a more precision heat cycle with the oven-within-an-oven concept.

The principle of the dual ovens is like the dual-door system that you find in the snow belt. The main door to the outside opens to a second door which is the main door for entry into the building or home. The idea here is that the main door does not open to the cold outside, but to a second door, minimiz-

ing heat loss in the winter months. The dual oven works on much the same principle: one heat coil to assist the other coil, for minimum temperature change.

TCXO Oscillators

A TCXO differs in circuit operation from an OCXO in that it does not have any oven-type control circuitry. How it controls the crystal vs. temperature is different. The TCXO circuitry uses variable components in connection with thermistors to make circuit adjustment as temperature varies. The process is much like a VFO and varactor capacitor

circuit. As the VFO now is varied, so is its frequency. In the TCXO oscillator the thermistor is much like the VFO in this application. The varistor varies its resistance, following temperature to make appropriate control voltage changes to maintain frequency accuracy vs. temperature changes. It's more complex circuitry, but it has many benefits over an OCXO. They are small-sized, and operate at low current. The cost for a TCXO runs about \$300 new; in surplus, something under \$50. For those interested: I have a purchased a limited quantity of these oscillators and will make them available; see the end

of this article for details.

The low current consumption of a TCXO can run around 50 mA. Circuits are temperature-rated over a large temperature range, usually over 100 degrees or more, depending on the quality of the oscillator. The oscillator shown in Photo A is the TCXO that I use in my portable 10 GHz station. Its accuracy over a large temperature variation is 1 PPM. Over a much narrower range of temperature (70 to 90 degrees F) or so, it is capable of much better stability, somewhat better than 0.1 PPM for short-term stability. Photo A shows a defective oscillator assembly removed from its

case. It's placed next to a complete unit. The TCXO is 2" square and 1/2" high. The defective oscillator shown had a shattered crystal, probably from rough handling. I eventually put it to use as part of a crystal test oscillator.

The stock TCXO uses three transistors, one voltage regulator, three thermistors, and one Johanson variable frequency-setting capacitor. The thermistors are thermally bonded to the crystal for maximum sensitivity to the crystal's temperature changes. The Johanson variable capacitor is quite expensive (about \$13 single quantity), and as such is a very accurate capacitor. It is

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part of the consideration and accuracy of this type of oscillator.

With regard to both the TXCO and the OCXOs just described, most units available are small compared to a soda-pop can. By comparison, an OCXO would be about half a can, and the TCXO would be 1/8th of a can in size. Moving into the arena of the next higher accuracy standard makes size comparison pale by contrast to a soda can. Frequency standards of higher accuracy can be small but are "UNOB-TAINUM" on the surplus markets. What do show up are the Frequency Electronics Incorporated (F.E.I.), Seltzer, and a few other types. Most are relay-rack mounted for 19" bays, and weigh in at about 15 to 20 pounds. The F.E.I. standard weight is 20 pounds, and has a small package profile 8" high, 4" wide, and 18" long.

The package is a self-contained oscillator with multiple oven control for the crystal oscillator. It has an on-board transformer-isolated regulated power supply, and a secondary power DC battery emergency operation. The emergency battery power is 24 volts at 2-1/2 amp capacity. The front panel has provisions for monitoring status of all the individual ovens and battery voltages in addition to oscillator output at various frequencies such as 100 kHz, 1 and 5 MHz. The part I was holding out on is that the accuracy of this "OVEN CONTROLLED BEHEMOTH" is 1 millihertz at 5 MHz. That equates to 0.0002

PPM, or a few parts in 10 to the 10th decimal place. Provision for calibration maintenance is provided by a 10-turn counter dial to adjust for portions of 10 to the 10th part adjustment for very accurate calibration.

For short-term stability it can be adjusted a little tighter in accuracy, but to accomplish this I would have to phase-track over a much longer time period, days instead of 10 to 12 hours. The effort involved and the benefits derived did not seem worth it at this time. I am not trying to go into rocket science or predict what will be an exact frequency with minimum error; gold-plating in this case is better left to the back of a watch than my frequency counter.

The battery operation in these standards is nice, especially when the AC power is removed. The batteries take over if there is nothing happening and operation continues for about six hours or so. A nice feature is the ability to use the batteries in a planned operation to assist in the calibration of other amateurs' frequency counters or standards. To do so, just unplug the standard (battery operation starts) and drive over to the other amateurs' locations. When you get there, re-plug into AC power to charge the batteries and take your time to recalibrate at this new location. As far as the F.E.I. standard is concerned, it never left AC power—as long as the batteries last, accuracy is maintained.

I tested this several times to see how well our "Traveling Clock" would

hold accuracy with battery operation, and was pleased to see that there was no change at all. Even when we experienced some harsh movement of the case, it tipped over in my station wagon on a trip to the grocery store. On the second trip it was treated a lot better—it got to ride on the rear seats with a seat belt holding it down. I only have one of these jewels and don't want to try the "DROP TEST" at this time.

Well, I hope this gives you some idea of what to expect in the line of frequency calibration and what type of accuracies to expect from your frequency counters and various oscillators you might happen to find in the surplus market. To obtain surplus standard oscillators, check out used broken frequency counters. A surplus military counter with a functioning high accuracy standard, even though "crunched," should allow you to at least remove the standard and discard the rest. Keep your eyes open, and good hunting.

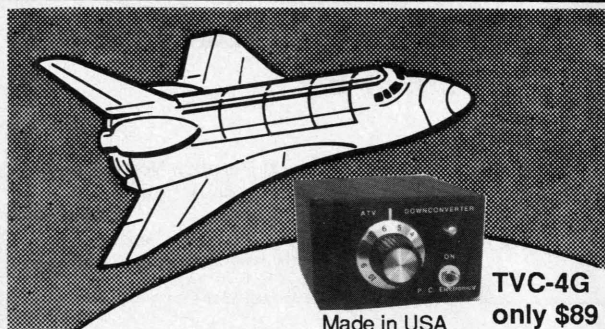
As always, I will be glad to answer questions concerning this and other aspects of our hobby. For a prompt reply please send an SASE. Concerning the



Photo B. OCXO oscillator: 10 MHz, oven controlled.

TXCO 10 MHz oscillators, I have a limited quantity available for \$40 each plus \$3 shipping (U.S.). I do run into a few higher accuracy types from time to time and will source them for you. Call for details. To order a TCXO oscillator send requests to Chuck Houghton, 6345 Badger Lake Ave., San Diego CA 92119. Well that's it for this month. Next month we'll get into simple microwave test equipment that can be home constructed. 73's Chuck WB6IGP. 73

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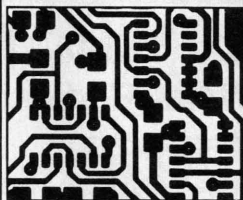
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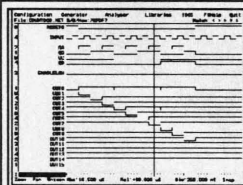
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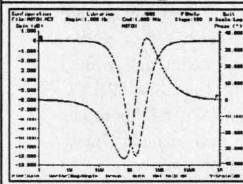


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Analog meters used as RF power output meters usually show relatively little average power output on SSB, especially as compared to keydown output. This often raises questions about proper equipment operation, especially when the owners' manual specifies "100 watts PEP" and the analog meter is reading considerably less. Too much mike gain and over-driving are often the result, with splatter across the operating band. The true power output during normal, everyday operation is often an unanswered question.

Measuring the peak and average power can be important. Two different rigs were returned to the dealer for replacement because on-the-air reports were of considerably less signal strength than comparable rigs, even though the key-down power was the rated value. Under a similar situation with a third radio, the problem was resolved by using a different type of mike on the radio. These

problems originated because the operator's voice frequency content and the radio audio processing were incompatible. Attempts to resolve these problems by measuring power levels with existing RF power meters proved unsuccessful because they were much too slow to follow the audio voice peaks.

The meter described here is designed to answer just such questions and to serve as a routine, everyday operating aid. It employs a remote sensor unit with a separate display that shows both average and peak RF output simultaneously. The output display unit can be used on the operating desk with the sensor unit out of sight under the operating position. The meter is left in the line and thus provides continuous indication that the transmitting equipment is operating normally. Abnormal readings may also indicate high SWR on the coax line.

The meter is easy to build, and uses readily available components. It is simple enough

to be built on perfboard. No special, tricky, or sensitive adjustments are required.

Photo A shows the assembled meter with an analog meter and a 10-LED bar display. Average output power is shown on the analog meter and peak power output is shown on the LEDs.

The Circuit

RF is sampled by a resistor divider and rectified in the remote sensing unit. The resulting signal is applied to two sections of an LM3900 op amp. One section of the LM3900 drives the analog meter and the other section drives a National Semiconductor NSM3916 module. The internal damping of the analog meter provides a good measure of the average RF power. The fast response of the LEDs follows voice peaks and provides a good indication of the PEP power. The LM3900 was selected because it operates easily from a single-ended power supply and was on hand in the parts box.

A 100-watt transmitter could drive the analog meter directly, but the op amp allows adaptation to other power levels if desired. In one case it allowed me to use an existing resistor divider located inside the cabinet of a transmitter (32 sheet-metal screws would have had to be removed and reinstalled to make a change to that resistor divider!)

One section of the LM3900 drives the 1 mA meter through a 10k calibration pot. The 10 LEDs are part of a National Semiconductor NSM3916 module purchased at a local hamfest. The second section of the LM3900 drives the NSM3916 module through a second 10k calibration pot. Note the 0.1 cap and 1M resistor connected to the NSM3916 module at pin 6. They provide the correct time constant for easy visibility of the LEDs as they follow the voice peaks.

Figures 1, 2 and 3 show the schematics. Figure 1 shows the display and control circuits; Figure 2 shows the RF sampler schematic. An alternate schematic for the NSM3916 module is shown in Figure 3. The values shown are those used with a transmitter with 100-watt output, and were selected

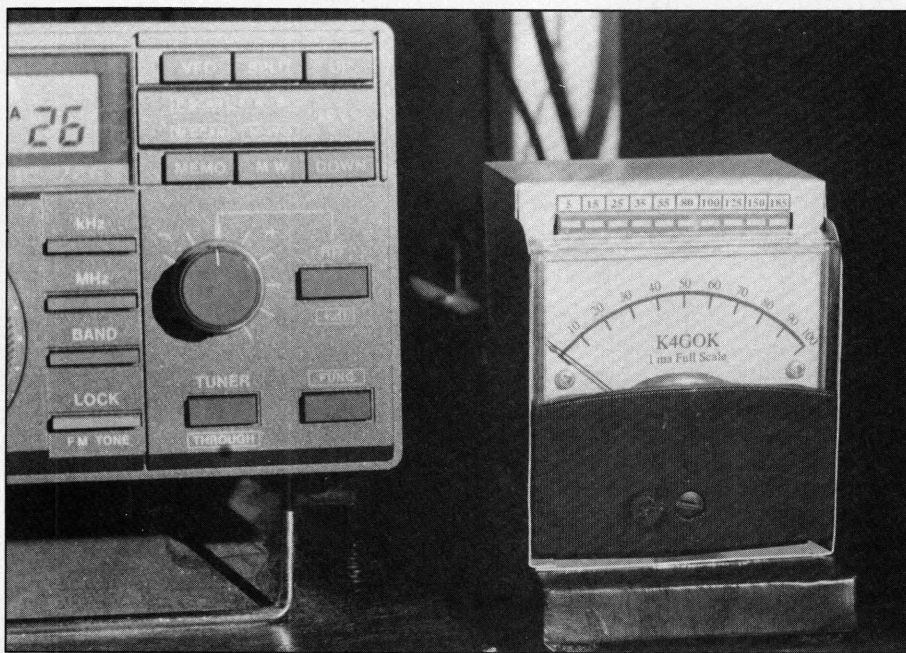


Photo A. The Peak and Average RF Power Meter.

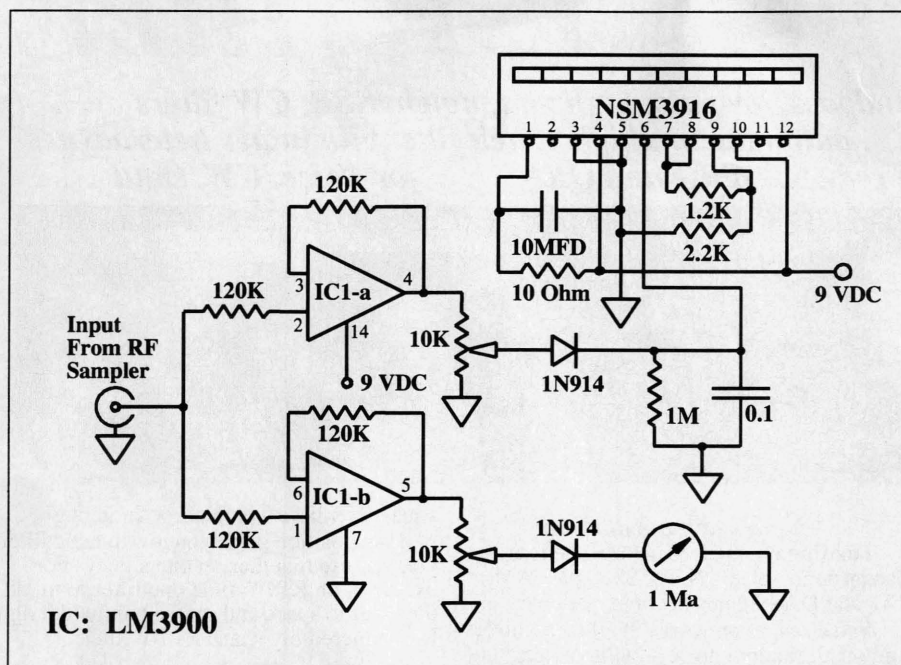


Figure 1. Schematic of the RF Power Meter.

because they were on hand and provided the desired performance. Other values could be substituted for the 120k resistors without problems, as long as both the input and feedback values are the same. The circuit as shown should work well with only a few watts of RF power by adjustment of the calibration pots and an appropriate RF resistor divider in the remote sensor unit. Don't hesitate to make changes in RF sensor box resistors to get the desired performance. And don't be surprised if the resistors used are frequency sensitive. The resistors shown in Figure 2 provide about 100 to 1 voltage division at DC, but about 10 to 1 at 50 MHz, suggesting that some of the "carbon composition" resistors are actually metal film types!

The RFC shown in Figure 2 should be selected for the band in which the power meter is to be used. Use 1 mH for HF, a Z50 for 6 meters, and a Z144 for 2 meters.

Construction

The recommended construction is simple and straightforward. It makes use of the NSM3916 module and perfboard. Fixed resistors are mounted directly on the module to fix the LED brightness and its voltage range. The other components are mounted on perfboard, except the RF sampling divider and RF rectifier diode, which are mounted in the remote sensor unit. The output from the sensor unit is fed to the indicator unit via RG 175 coax cable. The perfboard was mounted directly to the analog meter terminals. The NSM3916 module was mounted to the upper back of the analog meter using double-sided foam tape so that the LEDs are visible over the top of the analog meter.

Calibration

Calibration is best done with a known RF power source and an accurate RF power meter. A relative power scale can be obtained with a variable voltage source and a voltmeter. A Bird wattmeter can be used to measure the output of a transmitter at a known power level, say 100 watts, into a dummy load. (Use a dummy load to prevent QRM on the bands, please!)

First, set the pots to the low end of their range and apply 9 VDC or 12 VDC regulated power to the meter circuit. Next, apply the known RF power (key down) to the input and adjust the appropriate calibration pot for a full-scale reading on the analog meter. Then adjust the other calibration pot so that

the #7 LED lights. Switch to SSB and verify that LEDs #8, 9, and possibly 10, light on voice peaks. The calibration pot can be adjusted later to obtain the desired LED display on voice peaks. It is recommended that the adjustment be such that the #10 LED does not light under normal operation conditions so that it can serve to indicate overdriving, or abnormal operation.

Labeling each LED to indicate peak power levels in watts requires a variable RF source or controllable voltage source and a voltmeter. If you have an accurate wattmeter, you can slowly increase the transmitter output, note the power as each LED lights, and label them accordingly.

In the absence of an accurately known RF power source, the meter can be calibrated in a scale showing relative power levels. For example, I recommend that LED #7 be used as 100 watts, and the remaining LEDs be labeled in terms of power relative to that. Using a variable voltage source and voltmeter, connect the variable supply to the input of the display unit, in place of the input from the remote sensing unit. Set the 10k calibration pots to mid-range. Slowly increase the voltage and record the values at which each LED lights. When the #7 LED lights, adjust the correct calibration pot so that the analog meter reads full scale. Calculate the power level for each LED for each of the recorded voltages by using the formula below. First calculate "K" as follows:

$$K = 70.71/V_{\#7}$$

Then calculate the power for each LED:

$$\text{Power} = (K \times V_{\text{LED}})^2/50$$

You can make a label for the analog meter in a similar manner if desired. The analog meter shown in the photo was not calibrated with an RF power scale.

Conclusion

The peak and average reading RF power

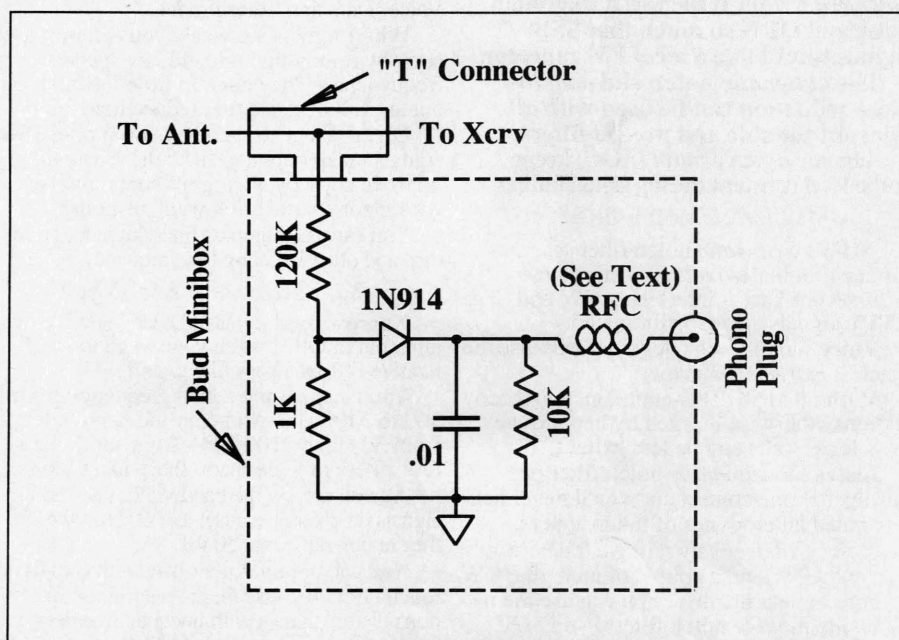


Figure 2. Schematic for the RF Sampler.

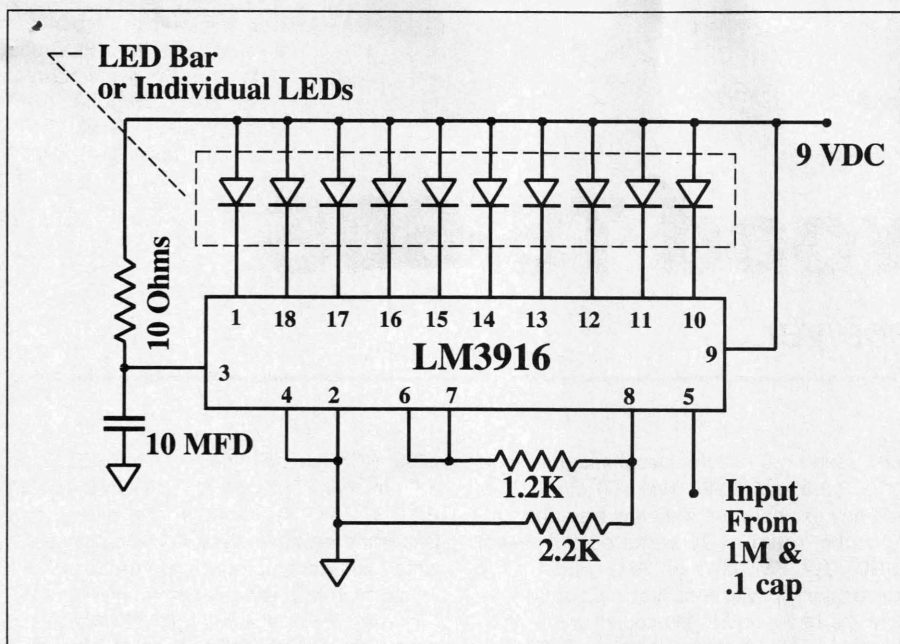


Figure 3. Schematic for an alternative to the NSM3916 module.

meter has been in use at this QTH for many years. It has been a valuable asset and a real "hassle-saver." It has proven reliable and a steady indicator that the solid-state "brick" I use on 6 meters is performing normally.

Readers are encouraged to build this simple power meter and enjoy the comfort of

knowing that "all is well" while operating.

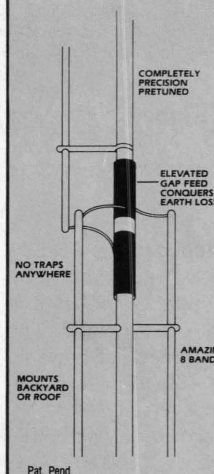
Drilled and etched PC boards for this project are available for \$4.50 plus \$1.50 S&H per order from Far Circuits, 18N640 Field Court, Dundee, IL 60118. Please specify either the NSM3916 or LM3916 module. Artwork for etching your own board is available by sending SASE to 73.

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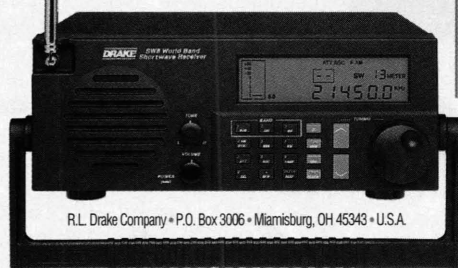
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Microwave Operation Tricks of the Trade

Well, tricks of the trade are not all secretive idioms, but rather they're just easy-to-use methods of procedure and equipment adjuncts to make microwave operation easier. For instance, when standing on a hilltop and trying to point your dish to a station on 10 GHz located some 50 to 100 miles from your location, where exactly do you point?

Also, how do you confirm that your station is operating on an exact frequency — let's say 10,368.1 MHz? Some preparation is needed before starting out to make hilltop contacts.

Additionally, to make remote contacts

you need a liaison frequency to first set up protocol of operation — like who is going to transmit and who is going to receive. What frequency is being used and such. Normally, most liaison communications are set up using either 2 or 3/4 meters.

Omnidirectional antennas or small beam antennas are normally used to begin setup of the communications dialogue. This is fine for setting up who and where to communicate with, but it does nothing on improving your receiver functions, frequency accuracy, and compass heading accuracy.

Where do you point your microwave dish antenna with certainty? Do you set up with a hand-held compass to align your tripod? Remember, most small microwave dish antennas have a 3-degree or less pointing angle, making pointing a dish antenna require accuracy in both vertical and horizontal angles. Larger dish antennas have smaller pointing angles of 2 degrees or less.

Well, the problem is not so difficult after all — it just requires some accessory equipment remotely located — namely a beacon transmitter to make pointing an easy function. A beacon transmitter remotely positioned in a good local spot provides a nearby signal to allow accurate pointing angles to be initially set up. Even if your location does not have a beacon transmitter, one can be set up on the way up to your hilltop spot by dropping off a portable source at a reasonable location for temporary use. Here in the San Diego area we have a local beacon set up keying CW identification followed by a solid key-down CW note for easy location and adjustment of exact heading direction. This makes a great receiver test and allows the exact positioning of a direct compass heading to be set up easily. If your beacon source is accurate in frequency, it also serves to calibrate your receiver with confidence as to frequency. If no one has confidence in frequency and they can copy the beacon, just move 100 kHz up from the beacon frequency and try this tack to make your contact. Several ways to acquire remote

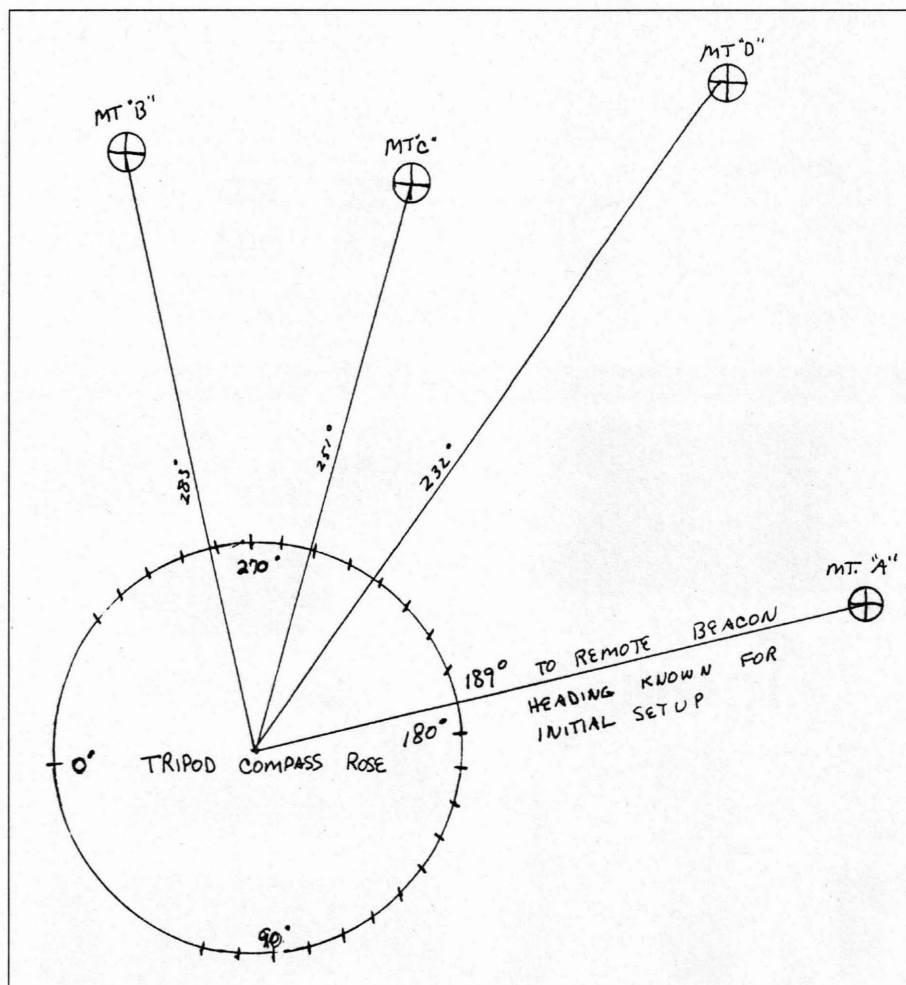


Fig. 1. Compass rose attached to tripod base enabling accurate alignment with a remote beacon, thus calibrating other microwave headings.

stations using microwave and dish antennas are possible.

Once aligned, the microwave dish antenna is properly focused on the remote beacon and the tripod and compass rose positioned to the heading in degrees from where you are to the remote beacon. All other locations are calibrated on the compass rose dial on the tripod mount. As long as the remote beacon's heading is known from your location and the compass rose is so calibrated by receiving a signal from the beacon, you then have one point calibrated. Also, all other points are calibrated by keeping the compass fixed in position and moving the tripod horizontally to the compass heading of the new direction where you wish to make test transmissions.

Knowing where you are set up and where the beacon is located removes the question as to where all other compass bearings (remote distant locations) are. These preparations can be worked out previously and logged on a simple lookup chart. Construction of the compass rose can be as simple as a paper compass on a firm backing, or even two 180-degree protractors forming calibrated 360-degree marks. In any case, whatever you use, mount this compass rose on your tripod below where the horizontal movement of the dish allows rotation of the dish antennas without movement of the compass rose.

Once you've done this simple procedure of calibrating your dish and compass heading to one distant heading, all other location headings are simple to set up. All you have to do is unlock your mount's tripod horizontal base, keeping the compass locked in direction to the beacon, and all other headings are easily aimed at by looking at the compass rose on your mount. All you have to have now is a list of possible locations remote to your location and what compass bearing they are to your location. It's just point and shoot.

Because the beacon is remote to your location — let's say a minimum of 5 miles distant — the vertical orientation will be in good agreement on stations located a hundred miles distant from your location. The difference between 5 miles and 100 miles due to Earth curvature is minimal and will be in good agreement with the beacon setup angles. Even if you want to calibrate this vertical angle it can be simply done by using an old meter movement and removing the spring on the meter vane so that the meter is free to swing like a pendulum. Mount the meter on your tripod upside down and now you have a calibrated vertical incline meter. All you have to do is reference the meter

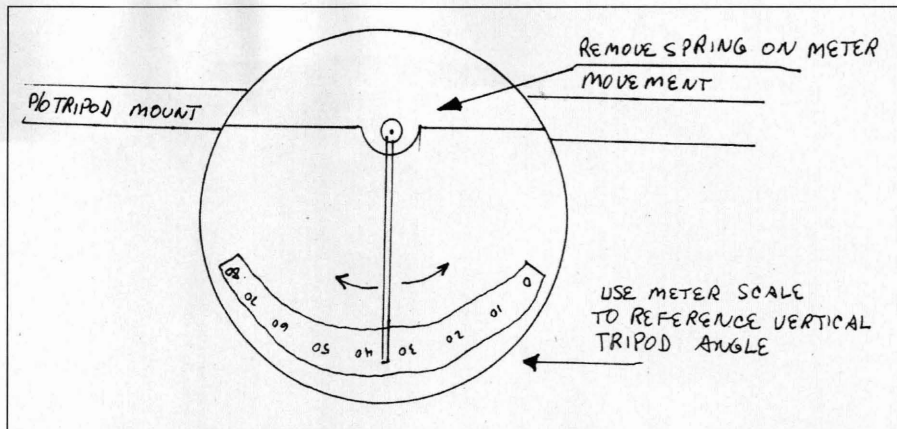


Fig. 2. Meter movement hung upside down (with meter movement spring removed). Becomes incline meter calibrating vertical dish movement.

calibration number from the meter dial (let's say 5.5 mA if it's an old current meter). This calibration is done when you are pointing and calibrated with the remote beacon. Recalibration can be determined at a glance by remembering this original calibration number (5.5 mA), which makes the operation like shooting fish in a barrel.

So much for the simple things. What if you don't have a beacon or frequency standard in your area? What can be done? Well, there are several things that can be done, none of which includes purchasing a beacon from commercial sources, as none exists at any inexpensive or modest price that I am aware of. What we can do is to construct a device home-brew fashion. There are several choices, whether a surplus low frequency brick or a synthesizer that normally operates

on a lower harmonic of the desired microwave output frequency. Finding something calibrated on the exact frequency desired would be a bonus shot if it could be located.

First, let's design a system for a frequency reference with an output of 10.368 GHz with excellent accuracy. Designing this source and constructing one from scratch would be a formidable task. Modifying one from surplus material is a far better success story. Originally, Kerry N6IZW, of the San Diego Microwave Group, constructed the first beacon here with a used Frequency West brick oscillator and a special-cut crystal in a TO-5 transistor-like case. This beacon is a great system and has operated for many years locally, providing a signal for receiver testing and antenna pointing heading alignment. While this device could be copied,

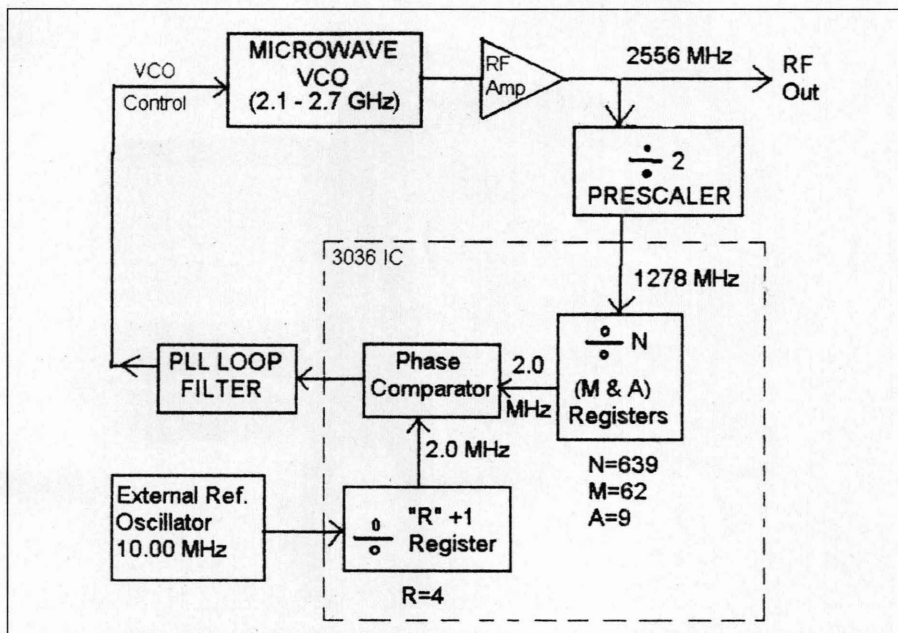


Fig. 3. Block diagram of Qualcomm synthesizer, Texas Version. Example shown for 2556 MHz and 2 MHz reference frequency. All components on PC board except 10 MHz TCXO oscillator.

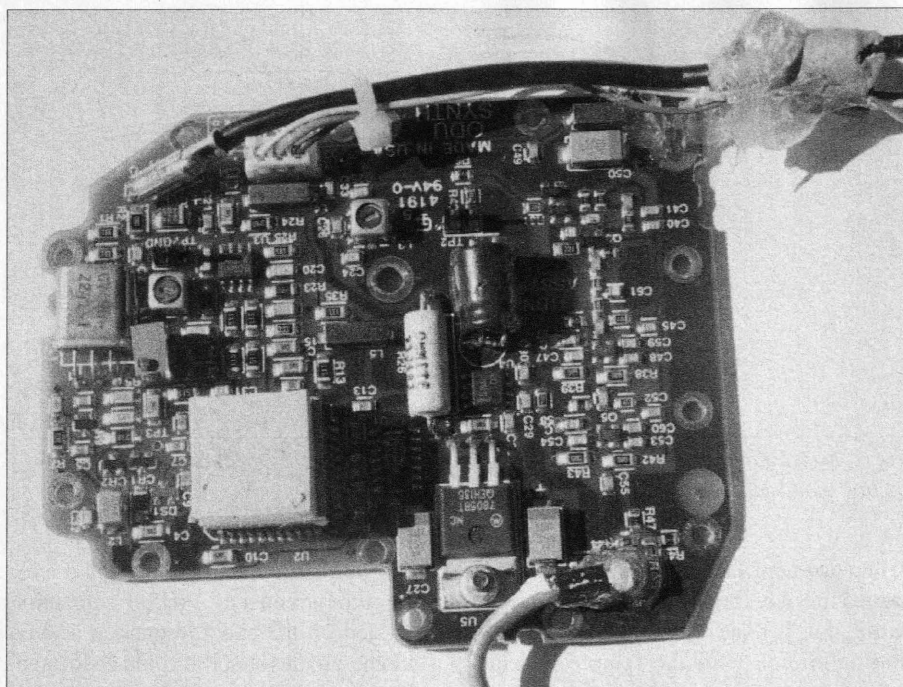


Photo A. The Texas Synthesizer. Main processor PLL chip visible as square white heat sink material. Called Texas board due to shape of unit.

there remain some material acquisition problems that the bricks of years ago present.

The problem with brick-type oscillators is the 100 MHz overtone crystal. It is getting hard to locate suppliers willing to fabricate an oven crystal at reasonable cost. It seems

the crystal required has fallen on low demand and industry wants larger-quantity orders. That makes the alternative plan to use surplus synthesizers more viable. Their attributes include being frequency-agile, small in size, and relatively inexpensive. Agility in frequency comes from the fact

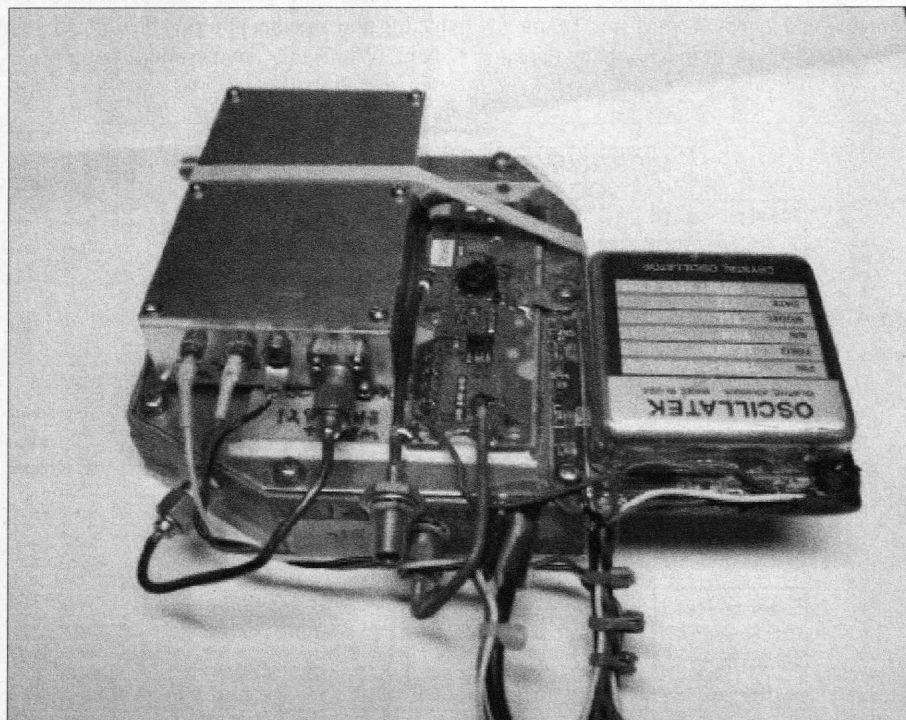


Photo B. Completed Texas Synthesizer mounted in heat sink metal compartment with 4x harmonic multiplier on top with another amplifier board for gain at divide-by-2 output. This unit can provide 1/2 LO output or 1,296 MHz, plus 2,592 MHz, multiply by 4 = 10,368 MHz. For band, very accurate band edge markers.

that they are driven by a standard frequency clock which is divided by a reference divider to a submultiple of the desired frequency. See **Fig. 3**. For a block diagram of a sample unit functioning at 2,556 MHz, which is the frequency for a 10,224 LO and an IF of 144 MHz = operation on 10,368 for a transverter. Another reason for using the synthesizers is that we have a modest quantity of the surplus material and will make the surplus material available.

Let's start out with the modification of a microwave synthesizer source to generate the 10,368 MHz marker signal at a subharmonic of 10,368 MHz. We have synthesizers that normally operate in the 2,300 to 2,800 MHz range in frequency steps of 2 or 5 MHz as determined in programming of the synthesizer reference divide-by counter. Dividing the 10,368 GHz operating frequency by 4 produces 2,592 MHz, which fills the bill nicely as the synthesizers run in the 2.6 GHz range normally. Modifying the original frequency from its 2,620 MHz is quite easy. It requires lifting several pins on the synthesizer chip to reprogram the divide-by counters to another value.

The synth is driven by a 10 MHz TCXO (temperature-controlled crystal oscillator) serving as master clock. For our application, the normal divide-by-2 (i.e., 5 MHz clock rate) is modified to divide by 5 to produce a 2 MHz reference frequency clock for the phase locked loop chip (PLL). This is necessary as the desired frequency in this application, 2,592 MHz, is not divisible by 5 but is divisible by 2.

The VCO (voltage-controlled oscillator) on this PLL board has a VCO oscillator that runs from about 2,300 MHz to just over 2,700 MHz normally without modification to the VCO. A divide-by-two chip interfaces the VCO and the synth chip, as the synth chip has a maximum frequency of 1.6 GHz maximum. The synthesizer we modified we call the Texas board due to its irregular shape — somewhat like the outline of Texas. These synths and clock oscillators are available from the author (see details at end of article).

To convert the Texas synth to its new frequency of operation of 2,592 MHz, lift with a sharp X-Acto knife pin #2 and short solder to pin #3 (Gnd). Lift pin #4 and leave open as it's pulled high (logic 1) by internal pull-up resistors. Lift pins 7, 8, 9, and 14, and leave high (open) logic "1". Lift pins 18, 19, and 21. Short pins 18 and 19 together to pin #20 (Gnd). Leave pin #21 open high (logic 1).

Continued on page 58

Meeting Your Match

The fine points of understanding matching networks.

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When we connect a load to a source, be it an antenna or an amplifier stage, we expect power to be delivered to the load. Whether the power delivered is the maximum available from the generator (source) or not depends on the resistance of the load relative to the internal resistance of the generator.

The maximum power is delivered to a conjugate matched load. A conjugate matched load has the same magnitude as the generator's internal impedance, but with the opposite phase angle. The reactive component of the load is then in resonance with the reactance of the generator. When the load impedance has the same magnitude and phase angle as the generator, the load is said to be matched on an image impedance basis. The term "image" arises from the fact that impedances on the two sides of the output terminals are images of each other. Only when the load and generator are purely resistive are the image match and the conjugate match the same.

In a voltage amplifier, say in a receiver IF, voltage is the important concern, not power. Therefore, the amplifier load impedance is maximized at the expense of power output. Under these conditions, matching is not important.

However, in a transmitter amplifier chain, power usually is critical and the amplifiers matched for maximum power output.

It's all well and good to say the maximum available power from a generator is obtained when the load is a conjugate match to the generator. If you designed the amplifier, you know what the impedance should be, but what is the output impedance of an unknown generator? The internal impedance of a generator can be found with a couple of measurements and some calculation. **Fig. 1** shows a generator with an unknown impedance. R_g is the generator's open circuit voltage divided by the short circuit current, that is, $R_g = E_g/I_g$. When the generator cannot be safely operated unloaded or shorted, the internal resistance can be calculated by noting the voltage across two different values of load. The reactance, either inductive or capacitive, is tuned out when the voltage across an arbitrary load is maximum. The voltage across the load is measured, the load is changed, and the voltage across the second load resistance is measured. The two different loads and load voltages produce two equations with two unknowns, E_g and R_g , which can be solved simultaneously.

For example, if the voltage across a 400 Ω load is 30 V and the voltage across a 200 Ω load is 20 V, the two equations for E_g are:

$$E_g = \frac{(EL1 + EL1R_g)}{RL1}$$

$$= \frac{(EL2 + EL2R_g)}{RL2}$$

Solving for R_g produces:

$$R_g = \frac{RL1RL2(EL1 - EL2)}{(EL1RL2 - EL2RL1)}$$

For the values of the loads and load voltages in the example, $R_g = 400 \Omega$ and $E_g = 60$ V.

Continued on page 50

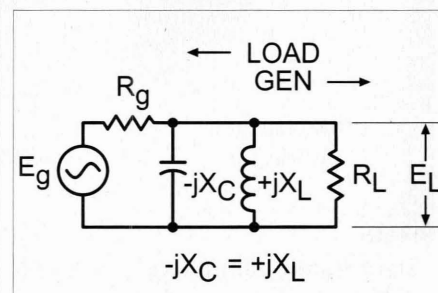


Fig. 1. The generator impedance can be calculated when the generator's reactance is "tuned out."

The relationship between load impedance, generator impedance, output voltage, and output power is shown in "Impedance and Power." As seen in **Table 1**, the maximum power is delivered to the load when $R_L = R_g$ and maximum voltage across the load occurs when the load impedance is maximum. To realize the desired load resistance requires an impedance transformer.

Transforming the load impedance can be accomplished with wound transformers or with suitable reactive networks. At low frequencies, the reactances required are not practical, and the shortcomings of wound transformers must be accepted. At RF, reactive matching networks are the norm. The impedance transformation of a wound transformer is fixed and determined by its construction—primarily the turns ratio. Therefore, wound transformers just transform the magnitude of the load impedance and seldom provide an image match or conjugate match.

The power output of a transformer or reactive network ideally is equal to the input power. Actually, the output power is a little less than the input power because of losses in the resistance of the windings and core losses. The equations for matching networks which follow assume lossless inductors and capacitors.

Matching networks made with discrete reactive components in the form of an "L", "π", or "T", or their combinations, are commonly used at RF. An "L" network is a "π" or "T" with one element reduced to zero or infinity. **Fig. 2a** shows a generic "π" network that matches the load R_o to the generator R_i . Usually the series arms and shunt arms have opposite signs. When the reactance Z_b in the "π" of **Fig. 2a** is infinite, the resulting network is an "L" as shown in **Fig. 2b**. **Fig. 2c** shows a "T" network. When the reactance Z_a in the "T" is zero, the resulting network is also an "L" as shown in **Fig. 2d**. Both the transformation ratio and phase shift through the network can both be selected in either a "π" or "T". In the "L", the phase shift is determined by the load resistance R_o and the series arm. The phase shift β through the "L" is $\tan^{-1}(Z_c/R_o)$, the angle whose tangent is Z_c/R_o .

The equations for the lossless reactances in an image matching "L" network are:

$$Z_1 = R_i \sqrt{[R_o / (R_i - R_o)]}$$

(input shunt arm)
[Equation 1]

and

$$Z_2 = \sqrt{[R_o (R_i - R_o)]}$$

(series arm)
[Equation 2]

The generator can feed either R_i or R_o , but R_o must be less than R_i .

In most impedance matching applications phase shift through the network is not important and the "L" network is satisfactory. But where phase shift is important, a full "π" or "T" may be required. The design equations for the "π" are:

$$Z_a = \frac{jR_1 R_2 \sin \beta}{[R_2 \cos \beta - \sqrt{(R_1 R_2)}]}$$

$$Z_b = \frac{jR_1 R_2 \sin \beta}{[R_1 \cos \beta - \sqrt{(R_1 R_2)}]}$$

$$Z_c = j \sin \beta \sqrt{(R_1 R_2)}$$

where β is the phase shift through the network. When β is 90° , $\sin \beta = 1$ and $\cos \beta = 0$, and the equations for the low-pass configuration reduce to

$$Z_a = -j \sqrt{(R_1 R_2)}$$

$$Z_b = -j \sqrt{(R_1 R_2)}$$

and

$$Z_c = j \sqrt{(R_1 R_2)}.$$

The network behaves like a quarter-wave transmission line transformer.

The design equations for the "T" are:

$$Z_a = \frac{-j[R_1 \cos \beta - \sqrt{(R_1 R_2)}]}{\sin \beta}$$

$$Z_b = \frac{-j[R_2 \cos \beta - \sqrt{(R_1 R_2)}]}{\sin \beta}$$

$$Z_c = \frac{j \sqrt{(R_1 R_2)}}{\sin \beta}$$

where β is the phase shift through the network.

When β is 90° , $\cos \beta = 0$, and $\sin \beta = 1$, the equations reduce to:

$$Z_a = -j \sqrt{(R_1 R_2)}$$

$$Z_b = -j \sqrt{(R_1 R_2)}$$

and

$$Z_c = +j \sqrt{(R_1 R_2)}.$$

Again, the network behaves as a quarter-wave transmission line.

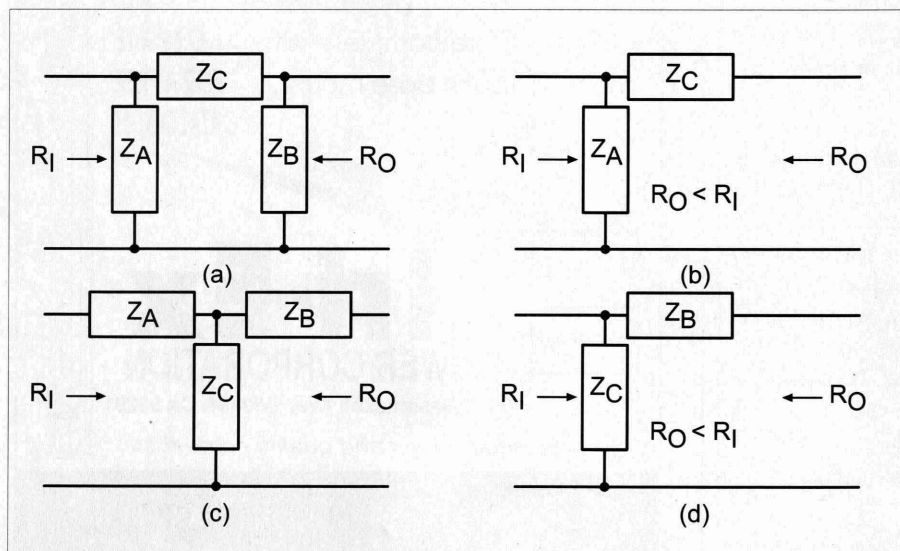


Fig. 2. There are many forms of matching networks. (a) A "π" network. (b) An "L" network derived from the "π." (c) A "T" network. (d) An "L" derived from the "T."

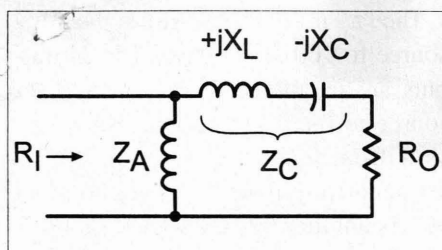


Fig. 3. A tuned circuit increases selectivity.

For an "L" to produce a resistance when looking into Z_a requires Z_a and Z_c to have opposite signs. Most often Z_c is a capacitor and Z_a an inductor, so that DC is blocked from the load and Z_a provides a series DC feedpath for the generator (plate or collector of an amplifier). Unfortunately, when Z_c is a capacitor, the "L" is a high-pass filter and there is no attenuation of harmonics. If harmonics must be suppressed, Z_c can be a series-tuned circuit operated off-resonance to provide the appropriate total reactance for Z_c .

The 3 dB bandwidth of a tuned circuit can be expressed as:

$$BW_{3dB} = \frac{F_o}{Q}$$

where

F_o = center frequency

and

$$Q = \frac{X_L}{R} = \frac{1}{2\pi F_o C R_o}$$

Making Z_c a series-tuned LC circuit as shown in Fig. 3 can produce a narrow

bandpass response and suppress harmonics. The net value for Z_c is $jX_L - jX_C$. The voltage across capacitor is equal to the current in the load times $-jX_C$. The current in the load I_L is

$$\sqrt{\frac{P_o}{R_o}}$$

Even with moderate power, the voltage across X_c can be surprisingly high.

For example, when you desire to transform 50 Ω to 800 Ω with a high pass "L", equations 1 and 2 show a series arm Z_c of $-j194$ and a shunt input arm of $+j207$. When you want to reduce the second harmonic, a series-tuned circuit operating below its resonant frequency has a capacitive reactance which can replace Z_c .

The response of a single tuned circuit falls 6 dB for every doubling of bandwidth. To achieve about 20 dB of harmonic suppression requires a Q slightly greater than 10. The reactance X_L then must be about $j500$ and the reactance X_c must be $-j694$ to produce the net reactance of $-j194$ required for the impedance transformation. When the power output is 200 W into 50 Ω , the current in the load and the series arm of the "L" is 2 A_{rms} or 2.828 A_{pk} while the voltage across the capacitor is 1.96 kV.

The network equations given in equations 1 and 2 produce a resistive input impedance, but a conjugate match will probably require an additional reactive component to tune out the generator's reactance. When the generator output has a capacitor to ground, a conjugate match requires an inductance in parallel to resonate the generator's capacitance. That additional inductance can be incorporated into Z_a .

For example, when the value of Z_a needed is $j207$ and the generator has a shunt capacity whose reactance is $-j1000$, the generator must be paralleled with $+j1000$ to tune out the generator's reactance. For $+j1000$ in parallel with $+jZ_a$ to be $+j207$ requires jZ_a to be $+j261$.

The "L" network is the simplest circuit for matching a load to a generator, and when the generator is a tube or transistor, the "L" can also provide DC

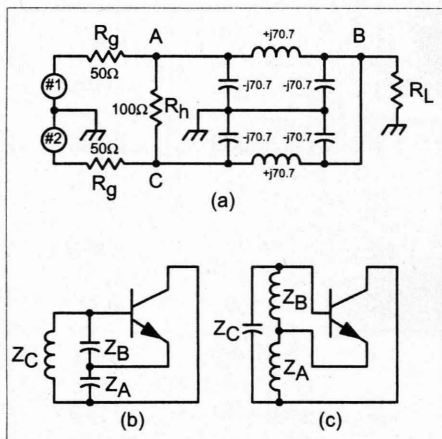


Fig. 4. Phase shift and impedance transformation are important in (a) a Wilkenson hybrid; (b) a Colpitts oscillator; and (c) a Hartley oscillator.

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Continued on page 52

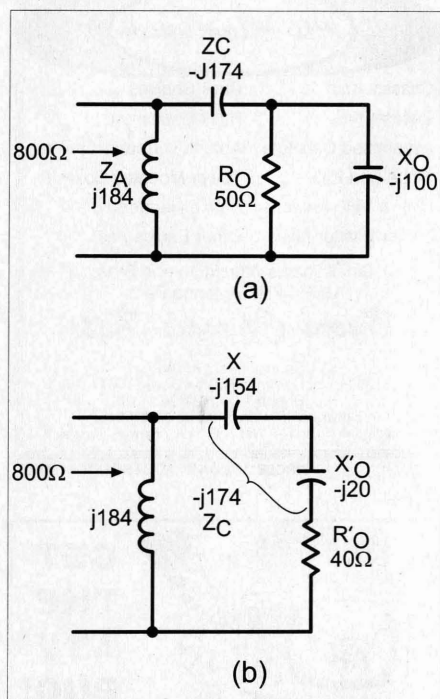


Fig. 5. An “L” can match reactive loads. (a) A load with a shunting capacitor. (b) The equivalent series circuit.

Meeting Your Match

continued from page 51

blocking to the load and a DC feedpath to the amplifier.

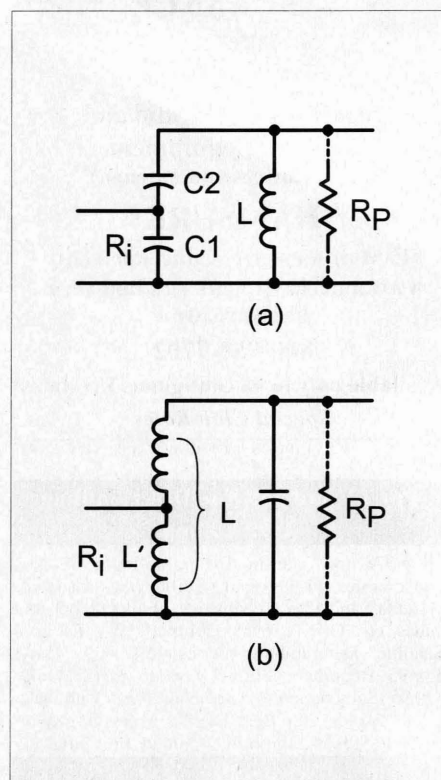


Fig. 6. (a) Tapped capacitors transform resistances. (b) Tapped inductors or an autotransformer can transform resistances.

The “π” matching network shown in **Fig. 2a** is commonly found as the plate or collector load of an amplifier. Z_b is often called the “loading” capacitor and Z_a the “tuning” capacitor. As in the “L”, here the selectivity can be increased by using a series-tuned circuit for Z_c .

Occasionally both the phase shift and impedance transformation are important characteristics of the matching network as in the Wilkinson hybrid shown in **Fig. 4**. Like any hybrid, the Wilkinson combines two inputs into one output and isolates the inputs from each other. Isolation means that input #1 does not appear at input #2 and vice versa. Without isolation, some of the signal #1 at point A will appear at point C and may modulate or pull signal source #2.

The “π” networks transform the 50 Ω source to 100 Ω. The two 100 Ω outputs are paralleled to present a 50 Ω source to the 50 Ω output. The phase shift from point “A” to point “B” is 90°, and from point “B” back to point “C” is another 90° for a total of 180°. Therefore, a non-phase shifted portion of input #1 applied to input #2 through R_h cancels #1’s signal coming back from the output. The two output capacitors can be combined into one with a reactance of 35.35.

In passing, the resonator of the Colpitts oscillator can be viewed as a “π” with 180° of phase shift. **Fig. 4b** shows the RF circuit of the Colpitts; the DC circuit is not shown. The input to the “π” is fed from the output of the amplifier and the output of the “π” is shifted 180° to drive the input of the

Impedance and Power

Fig. 7 shows the equivalent circuit of a generator and load. When the load is a resistor R_L , the power dissipated in the load is $I_L^2 R_L$ or E_L^2 / R_L , where I_L is the current in the load and E_L is the voltage across the load. The generator is matched when R_L equals R_g , and the voltage across a matched load is $E_g / 2$. When the load is an impedance Z_L , the voltage across the load is:

$$E_L = \frac{E_g Z_L}{(Z_L + R_g)}$$

but only the real (resistive) part of Z_L dissipates power; the reactive part stores energy, then gives it back. The resistive part of an impedance Z_L is:

$$R = Z_L \cos \theta$$

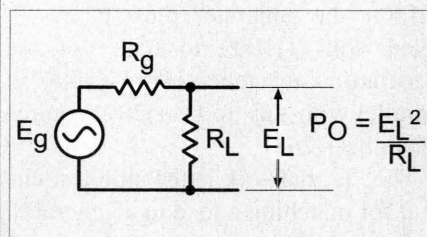


Fig. 7. The power output available from a generator depends on the load resistance compared to the generator.

The relationship of power in the load to the generator’s internal resistance is given in **Table 1**. From the values in the table, you can see that maximum power is obtained when the load resistance is equal to the generator resistance.

R_L / R_g	E_L	P_L
6	1.714	0.490
5	1.666	0.555
4	1.60	0.640
3	1.50	0.75
2	1.333	0.888
1	1	1.0
1/2	0.667	0.888
1/3	0.5	0.75
1/4	0.400	0.64
1/5	0.333	0.555
1/6	0.284	0.490

Table 1. Relationship of power in the load to the generator’s internal resistance.

amplifier. If the shunt arms of the “ π ” are inductors and the series arm a capacitor, the oscillator is seen to be the Hartley oscillator shown in **Fig. 4c**.

When the load R_o is in parallel with a capacitive reactance $-jX_o$, as shown in **Fig. 5a**, the equivalent series circuit is shown in **Fig. 5b**. X_o and R_o are transformed to their equivalent series components X_o' and R_o' . 50 Ω in parallel with $-j100$ is equivalent to 40 Ω in series with $-j20$.

“Conversions” describes the process of converting parallel and series Rs and jXs to their series equivalents. These calculations are rather time-consuming with a four-function calculator, but tolerable with a “scientific” calculator that has trig functions, squares, and square roots. A calculator that can compute polar and rectangular coordinates makes the calculations a snap.

For example, when $R_o = 50$ and $X_o = -j100$, the equivalent series circuit shown in **Fig. 5b** is $R_o' = 40$ and $X_o' = -j20$. The “L” network that transforms 40 Ω to 800 Ω is found with equations 1 and 2 to have $Z_c = -j174$ and $Z_A = +j184$. Since X_o' will be part of Z_c , the series arm of the network Z_c is:

$$Z_c = -jX_1 - jX_o' = -j174$$

When $-jX_o'$ is $-j20$, then $-jX_1 = -j154$. The final “L” has $Z_c = -j154$ and $Z_A = +j184$.

Fig. 6a is **Fig. 5** redrawn as the tapped capacitor tuned circuit often used in receivers to match the antenna to the input and reduce the loading on the input circuit. R_i is the input resistance, the antenna resistance, and R_p the transformed output resistance. Near resonance, the effect is to make the impedance offered to the input terminals less in magnitude than the parallel impedance of the entire circuit without changing the character of the impedance curve as far as shape or equivalent Q is concerned. Tapping a parallel resonant circuit accordingly offers a means of adjusting the magnitude of impedance obtained without changing the characteristics of the circuit itself. The ratio of impedance offered by the input R_i to the parallel impedance of the circuit R_p is:

$$\frac{R_i}{R_p} = \left[\frac{C_1}{(C_1 + C_2)} \right]^2$$

For example, if R_i is 50 and X_1 is $-j5$, then their equivalent series values are $R_i' = 0.49$ and $X_1' = j4.9$. If X_2 is $-j50$, then the total capacitive reactance across L is $-j55$. The equivalent Q is:

$$Q = \frac{X}{R} = \frac{55}{0.49} \approx 112$$

A tapped inductor follows the same procedure as the tapped capacitor network when there is no mutual coupling between the inductors. But, when there is mutual coupling as shown in **Fig. 6b**, $R_i/R_p = (M_{eq}/L)^2$, where M_{eq} is the total equivalent mutual impedance between L' and the entire coil L, including both common and inductive coupling.

Continued on page 56

Conversions

The impedance of the series circuit shown in **Fig. 8** is the vector sum of a resistance R and a reactance jX. The voltage across a resistor is in phase with the current in it and is shown on the horizontal axis. The voltage across a reactance is 90° out of phase with the current in it and is shown on the vertical axis. In an inductor, the voltage leads the current by 90°, and in a capacitor the voltage lags the current by 90°. The voltage across an inductor is expressed as jX_L , while the voltage across a capacitor is $-jX_C$. The factor j rotates a vector 90° in a counterclockwise direction. **Fig. 9** shows the vector sum of R and $+jX$. The resistive voltage is on the horizontal axis and the reactive voltage on the vertical axis. Their sum Z is the vector addition of R and $+jX$. From the Pythagorean theorem,

$$Z = \sqrt{(R^2 + X^2)}$$

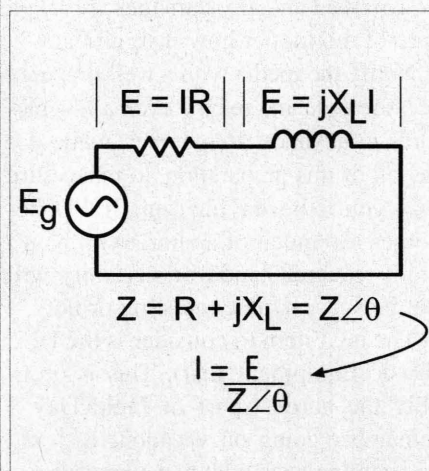


Fig. 8. A resistor and inductor in series add vectorially.

and from trigonometry, the angle θ is the angle whose tangent is X/R ($\tan^{-1} X/R$). Also from trigonometry, $X = Z \sin \theta$ and $R = Z \cos \theta$.

The horizontal axis represents the real component, resistance R or conductance G ($G = 1/R$), while the vertical axis represents the imaginary component, the reactance X or susceptance B ($-jB = 1/jX$).

$R_1 + jX_1$ can be added to $R_2 + jX_2$. Reals are added to reals and imaginaries added to imaginaries: $(R_1 + R_2) + (jX_1 + jX_2)$. Of course, $R_1 + jX_1$ can be multiplied with $R_2 + jX_2$, but it's much easier to do so in polar form:

$$Z_1 \angle \theta (Z_2 \angle \phi) = Z_1 Z_2 \angle (\theta + \phi)$$

Treat the angles as exponents.

Dividing rectangular forms is also much easier in polar form:

$$\frac{Z_1 \angle \theta}{Z_2 \angle \phi} = \left(\frac{Z_1}{Z_2} \right) \angle (\theta - \phi)$$

Again, treat the angles as exponents.

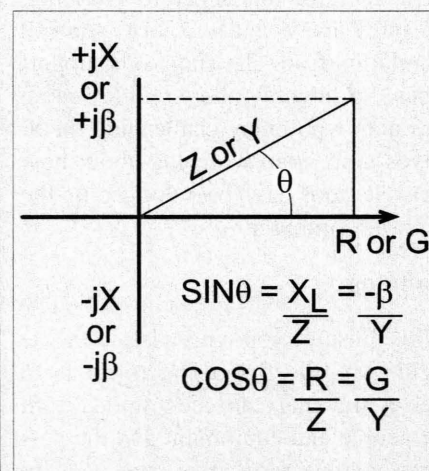


Fig. 9. A circuit containing R and X is depicted as a right triangle.

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In keeping with the amateur tradition of using the cheapest means possible to accomplish results, I have assembled complete 1.2GHz receive systems from old satellite receive equipment. The desired satellite receivers are known as block downconverted receivers. In their original implementation, they act as a tunable IF strip. The frequencies that they are designed to tune are 950MHz to 1450MHz. This makes them excellent candidates for use as receivers in our 1.2GHz FM ATV band. The front end of these receivers is very broad, so simply attaching an antenna to them usually gives poor results—you're weighed down with reception of broadcast TV signals, paging transmitters, cell phone intermod, and other signals. Also, they aren't usually very sensitive, so you have to add some type of preamp to bring the received signal up to the input level the receiver wants to see.

The first problem, that of the undesired signals, is cured by using a waveguide transition made from coffee cans as an antenna. The design for this feedhorn is available in various books and magazines, but it's simple enough to

cut the horn feed off and attach a type-N connector are okay if you've been around microwave equipment for a while, but I think it's time to give a more detailed description so that others can enjoy the kind of fun we're having here in Houston.

"We just ask homeowners with incorrectly-aimed dishes if we can have the LNA."

run through right here. Solder two 34.5-ounce coffee cans together so that only one end remains sealed. Place a type-N connector with a two-inch #14 wire soldered to the center pin three inches from the sealed end. You now have a 10dB "cantenna" to use in front of your receiver.

The hardest part of this project was the preamp. High-gain, low-noise preamps for use at 1.2GHz are not cheap. I found a solution in an old copy of 73. C. L. Houghton WB6IGP, in his "Above and

Beyond" column (May 1993), described a conversion of old satellite LNAs (low noise amplifiers) for use as preamps on spectrum analyzers. These LNAs were used on the earlier-style TVRO satellite systems and they were designed to pass 3.7 to 4.2GHz. His instructions to

First, find an LNA

The LNA is easily identified by the type-N connector on one end. We have a lot of luck finding these LNAs around Houston: We just ask homeowners with incorrectly-aimed dishes if we can have the LNA, if they aren't using the dish (see **Photo A**). Sometimes the owner says yes, but only if we take the dish—which is fine—these dishes perform well at 1.2GHz. I have a six-footer on my van, with a cantenna feedhorn, which has worked our repeater from numerous special events locations.

Place the LNA in a vise and carefully cut off the waveguide input section, leaving an area to mount the type-N connector (see **Photo B**). Remove the screws which hold the cover on the preamp. If a circulator/isolator is in your preamp, it must be removed. Look around inside the preamp to see how the unit was originally assembled. It may be necessary to remove all the screws and unsolder and remove the type-N connector to get the isolator out. The isolator must be removed as it is a narrowband device and we are moving

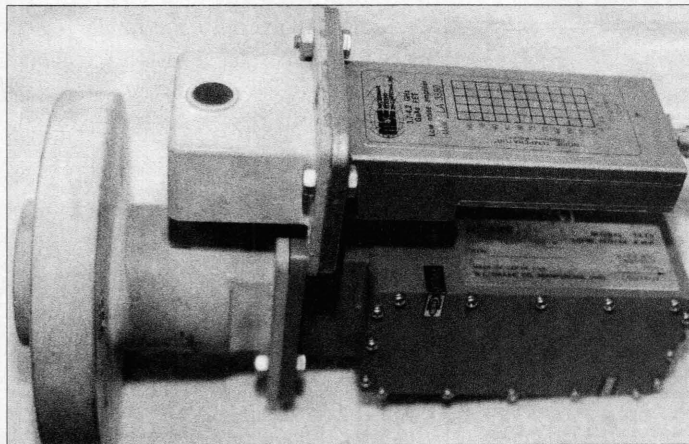


Photo A. Two LNAs and a two-axis feedhorn, hot off a satellite dish.

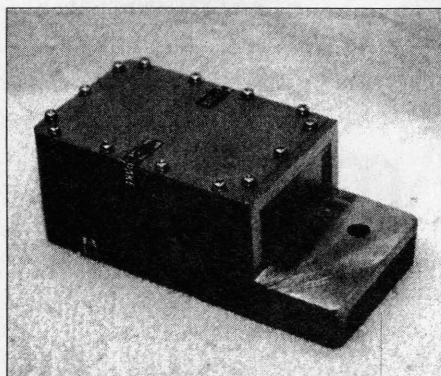


Photo B. Waveguide has been sawed off.

the designated passband down quite a bit (see **Photos C and D**). With the isolator removed, reassemble the preamp and mount a type-N connector near or in the hole where the original 3.7GHz waveguide probe was. Use either mini hardline or double-shielded mini coax to connect the type-N connector to the preamp's input. When you place the connector and its mounting be sure it's sealed watertight, because the LNA was designed to be mounted outdoors, and it'll take a beating from the elements. Both the center conductor and the shields of the coax must be attached to prevent oscillation of this very high-gain device. These units have an advertised gain of 40dB, and we have measured one device which had 60dB of gain with a 2.5dB noise figure after modification.

At the other end of the preamp, a comb filter may be included to limit the passband of the preamp. Removal of the filter is accomplished with a hobby knife. Only the tips are left, and a wire jumper is installed across them to try to keep stripline integrity (see **Photos E and F**).

You tweak the amp (if necessary) by removing or relocating blocking capacitors on the striplines. Older-type amps

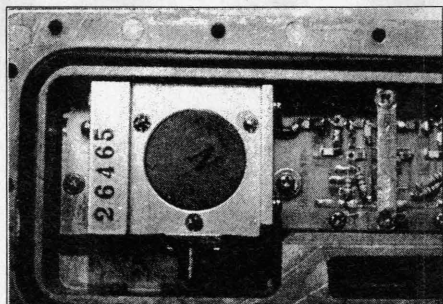


Photo C. Circulator to be removed.

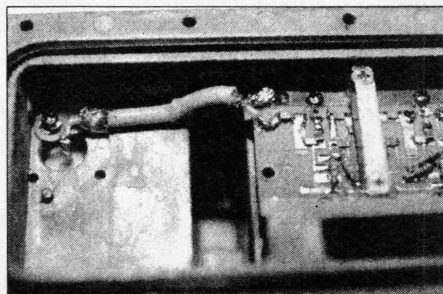


Photo D. Circulator removed and coax installed.

have two blocking caps on the microstriplines which decouple the stages in the preamp. Removal of the capacitor nearest the GaAsFET electrically lengthens the microstrip, lowering the operating frequency. Be careful to prevent oscillation, as these are very low-noise, high-gain GaAsFETs in these amps. Generally, amps with five or more active devices should not be tweaked. Amps with three or four stages can be heavily massaged to lower their operating frequencies. Coupling capacitors between stages can be increased to aid low frequency response. Usually, while tweaking, I will simply add a capacitor beside the original in case removal is required later. Tweaking is a trial-and-error approach, and should not be a substitute for proper preamp design.

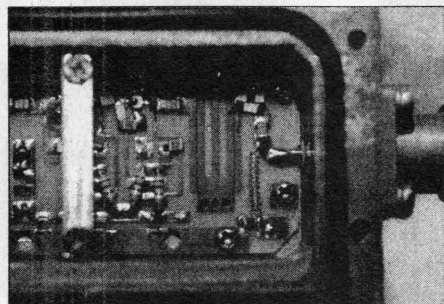


Photo E. Comb filter to be removed.

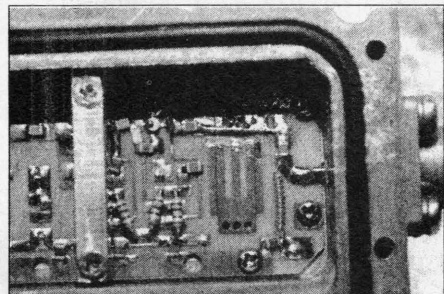


Photo F. Comb filter bypassed with wire jumper.

When completed, the modified preamp is a gain-block type of device. It should be very broadband, very sensitive, and fairly quiet. Use on other bands or modes is possible—Phase 3D comes to mind. The same requirements for frequency selectivity mentioned about the satellite receivers apply to these LNAs. Some type of frequency-selective filtration must be applied in front of the amp. Use of a commercial 1.2GHz yagi directly into the preamps was unsatisfactory.

We've had some success in Houston with a two-cavity filter which allows full duplex operation on 1.2GHz. Separate antennas and careful placement are required. We transmit on 1.255GHz and receive on 1.285GHz. The use of these satellite receivers, only 30MHz removed from 18W transmitters, has proven them a practical substitute for much higher-priced commercial FM ATV receiver systems. 73

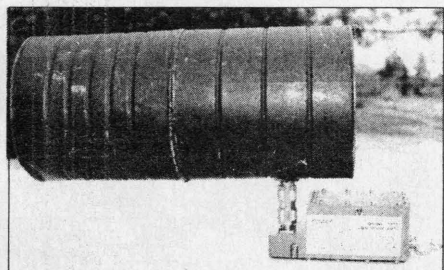


Photo G. Antenna and preamp ready for service.



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Log Amplifiers

This month I want to cover a commercial specialty amplifier called a log amplifier. Last month I covered a homebrew 30 MHz RF preamplifier and some component selection considerations. The preamplifier was intended for use in RF applications for power and low noise receiver or transmitter applications. The log amplifier is similar to the RF type amplifier, with some circuitry changes allowing its output to be handled in a logarithmic fashion. Let's explore this special type of amplifier and cover some amateur radio applications.

A log amp is a multi-stage IF amplifier that has some special circuitry added by design into the amplifier. Basically, the circuitry consists of a diode summing network and output stage. The log amplifier has eight or so stages, with each stage's gain/dynamic range set to 10 dB. The output of each one of these eight or so stages is connected to a diode summing network/limiter. The outputs of these diodes are all tied by common feeding of a video output or associated video amplifier. The output of this video amplifier is the business end of the log amplifier circuit. Additional connections may be made to bring out the standard IF output on the log amplifier, labeled "linear" output. All log amps can be coupled with this output, but not all are.

This makes the amp very usable in both a log function and as a common IF amplifier for a multitude of projects. If used in this case (as an IF amp), it will have a gain of some 70 to 80 dB gain at the amp's specified frequency. Figure 1 shows a block diagram of a typical log amplifier. Most amplifiers have a preamplifier and a buffer amplifier used to drive six to eight identical stages in tandem.

Each stage is set for about 10 dB of dynamic range/gain as determined by circuit parameters.

Trawling for Log Amps

Locating a log amp in surplus can be somewhat misleading. The ones that I have located came disguised inside plug-in type units with shielded cover plates. I suppose they were made to be inserted into a mother board assembly. They look unobtrusive on the outside, having lots of inputs and output connectors, including several SMA types. In other words, they look like something you might not want to pick up unless you are into scrap aluminum. Looking at the designations on the connectors can help you determine what might reside inside, if it is labeled. Most I have found were not labeled with anything I could understand, just cable hookup designators for a specific piece of equipment.

Details to look for are amplifier video output, switched IF in, linear or, even better, log out. These are some of the key words used. Photo C shows an amplifier stating 0.75 to 18 GHz and connectors labeled log, lin, RF in, and power. It does not go to 18 GHz as stated, but was part of a radar receiver assembly that did go to 18 GHz. This amp is strictly low frequency RF, 60 MHz and lots of gain (90 dB).

If you are confused or not sure about an amplifier you might locate, take the cover off to see what is inside. (That's why you should take a small tool kit to a swap meet). Sometimes it can be discrete circuitry and not of importance except for parts. However, if you don't look, your chances of finding a log amp in surplus are slim to none.

There are many different manufacturers who produce these devices in ranges that cover frequencies from a few MHz to about 1 GHz or so. Commercially, they cost upwards of \$1,000 each. This cost is indicative of miniature

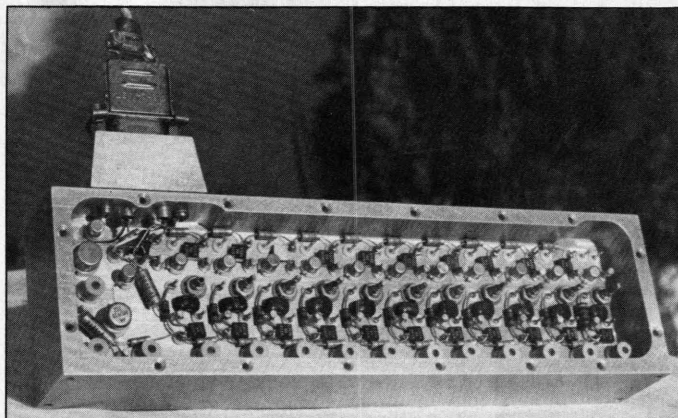


Photo A. Old discrete 60 MHz log amplifier, 12" long. This one has 11 stages.

monolithic substrate type of construction, with assembly under a microscope necessary. Accordingly, they are very small; several can fit into a cigarette flip-top box. Photo B is typical of this type amp. Commercial cost for this baby is \$1,700. How many do you want?

I always keep my eyes open at our local electronics swap meet. Some time ago I ran into a dealer having a box of seven or eight modules that looked quite nondescript on the outside. They were inexpensive (less than \$10 each), so I looked under the cover. Bingo! I found not just an amplifier, but a log amplifier and several SMA .141 solid coaxial cables.

What to Look For

Now you say, "What is a log amplifier?" Well, it's not a preamp, but it is quite remarkable. It is an instrument amplifier that has a bandwidth centered about a design center frequency, say 60 MHz. The log amp has multi-stages that are all cumulative, summed in a diode network to form a video output of the total amplifier string. The issue isn't the gain of this type amplifier; rather, it is the dynamic range which relates to input sensitivity (in dB) that will still meet output linearity requirements. In other words, we want minimum sensitivity that still can produce a true reproduction at its output and keep the output relationship different for strong signals in the presence of weak signals.

Quite an amplifier.

Uses for a log amp can range from IF amplifiers for spectrum analyzer applications to electronic countermeasures receivers for military applications. It's the ability of this log amp to be able to qualitatively analyze low-level signals in the presence of high power signals that makes it special. In this application it keeps the output representation of these signals in their original condition, (linear). The output is presented as a video signal that varies between 0 and 2 volts; 2 volts represents the maximum signal received.

A typical log amp has eight or so stages of gain, with an impedance matching preamp and frequency filter in the first stage. This first stage sets the frequency bandwidth and VSWR of the unit. The next seven or eight stages comprise the main IF amplifier. The summing diode/limiters are all in common with a video output op amp for video information output signals.

Building a Spectrum Analyzer

Putting these devices to work in constructing a spectrum analyzer is quite easy compared to constructing one by hand. The components needed to accomplish this are: an oscilloscope for display; a voltage-controlled oscillator (VCO); sweep control circuits to produce a DC ramped sweep voltage to drive the VCO; and a bandpass filter and log amplifier (see Figure 3). The

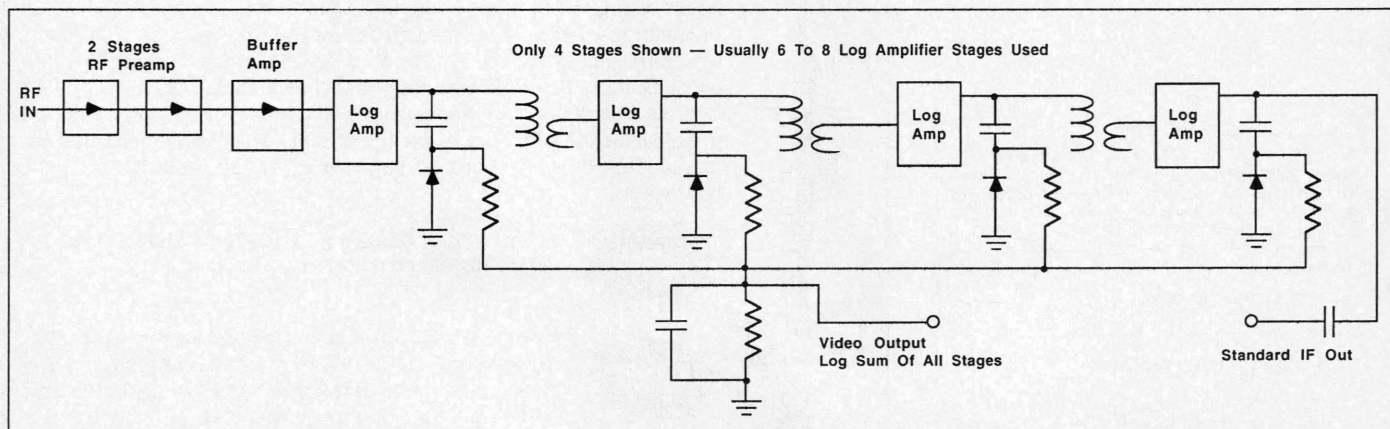


Figure 1. Block diagram of a typical log amplifier. The input preamp is usually followed by a buffer amp stage. Other stages cascade in normal RF fashion, but note the diode coupling network. It is a summing network for video information output. Each stage adds voltage to sum total of output voltages in logarithmic order. This can be used as a standard IF amp if the connection to the last amp stage has output from LC tank circuit.

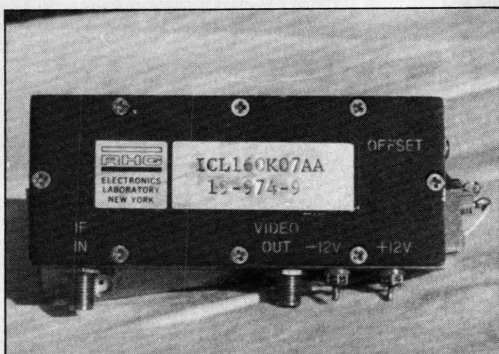


Photo B. New monolithic 160 MHz log amp 2-1/2" long. This one has 90 dB gain with 60 MHz dynamic range.

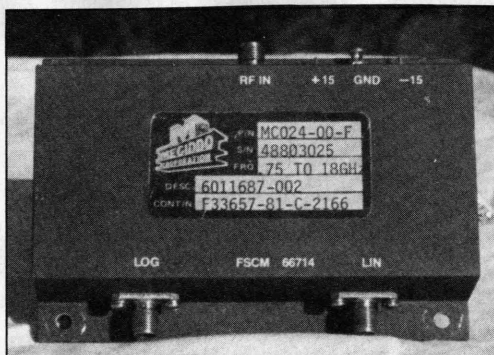


Photo C. Monolithic 60 MHz log amplifier has both video and linear (normal) IF output. Gain is 85 dB at IF, -90 dB for logarithmic (linear) applications.

main consideration for bench use or mobile use is dictated by the o-scope requirements. If you have a small pocket o-scope, the unit will be, needless to say, a versatile base or mobile. If you have to drag an AC plug for the scope, a very long extension cord may be required.

An alternative for the o-scope in surplus would be a medical heart-rate o-scope display unit. These are the units you see on many different emergency-type TV shows that show a scope representation of your heart rate on the screen. Sometimes you hear them "beeping" along with the heart rate. A straight line on one of these babies is bad news, to say the least. For our spectrum analyzer application, they are ideal. They have long persistence not available on normal o-scopes, allowing the trace to remain intensified on the scope screen, with slow sweep cycles.

Being very solid-state and rather new, most operate from AC-derived +12 volts DC power (what a break for us—+12 volts, an easy modification). Also, they are already set up with an internal sweep providing a DC ramp which can be brought out for our VCO; thus we don't need to construct that part

of the circuit. Connect an RF preamp and mixer with the VCO and a medical scope and of course our log amp and you have a spectrum analyzer. True, it's not an HP unit, but it does work.

I am not going to go into the mixer or RF circuit here except to say that a suitable unit could be constructed from a TV solid-state tuner. This module covers both the VHF and UHF TV frequencies, and has a built-in oscillator mixer and VCO oscillator IF amplifier. The normal IF output is quite broad and can be anything from 30 to about 90 MHz in most units, which is ideal for this application. Couple an IF filter and the log amp with a suitable display and you have a spectrum analyzer that can be used from about 50 to 800 MHz. This of course has some frequency gaps due to the standard TV frequency assignments, but it does give very good coverage.

I wish I had this approach available to me years ago. Back then I saw an article in the August 1982 issue of 73, titled "Poor Man's Spectrum Analyzer." This article gave full construction details on assembly, including details on construction of a log amp. I not only built the unit but designed PC boards for the

project and built the entire unit. It worked fine, covering 10 MHz to over 250 MHz. This was my first spectrum analyzer, of many that were to follow. The log amp in that case was constructed out of 40673 dual-gate MOSFETs for each stage in the log amp. The IF transformers were hand-wound toroids from Amidon and resonated at 60 MHz. Did it work? You bet, but the log amp was about 10 inches long and about two inches wide. Not too bad for home construction. I still have this amplifier around, along with many other early projects, to remind me of my humble beginnings.

Now, if construction of that log amp back then doesn't sound too humbling, maybe I should tell you that the first PC board I ever built used house paint as a resist, applied with a small artist's paint brush. Crude but it worked. I guess I am trying to make my point the hard way. It all boils down to this: You do not need to have a computer to do your design as it can be done with pencil and paper. All the computer does is do the same job faster. If the original thought is not put into proper perspective, there is no fast computer that can help.

Back to the home-built log amp us-

ing 40673 FETs. Initially I saw the original article as an interesting project, but without PC boards it would be too difficult to construct. After some time it was determined that I would construct PC boards in order to complete this project. I laid out the component parts on paper using a set of drawing aids that mirror component part sizes either in 1:1 size or two times scale for drawing ease. After laying all component parts on the paper from the schematic, it was just a matter of connecting all the dots from each component to make all connections.

Some of the components may have to be moved to make connections but it's kind of easy once you get the hang of parts placement. It's about as difficult as working a familiar crossword puzzle and many times the fun. Check out Photo D, showing the original spectrum analyzer that I constructed. It works nearly as well as its commercial counterparts. If you can't locate one in surplus (miniature unit) you can build a log amp just as I did. I still have the artwork and silkscreens available and can reproduce the PC boards if there is interest.

By today's standards it's quite large for its function, but it still can hold its own when stacked up to a commercial equivalent. It was made for 90 MHz and has 80 dB of gain and exhibits just over 75 dB of dynamic range. It was constructed over 10 years ago when the alternative available to me was the converted (tube type) IF strips out of WWII radar receivers.

At that time the miniature log amps used commercially were truly UNOBTAINIUM! Most of the test equipment and station apparatus were tube-type equipment. How our lives have changed! Back then my spare tube stock would rival a refrigerator box for space; today my spare tube stock consists mostly of museum pieces.

Whatever type of log amplifier you acquire or build in your application,

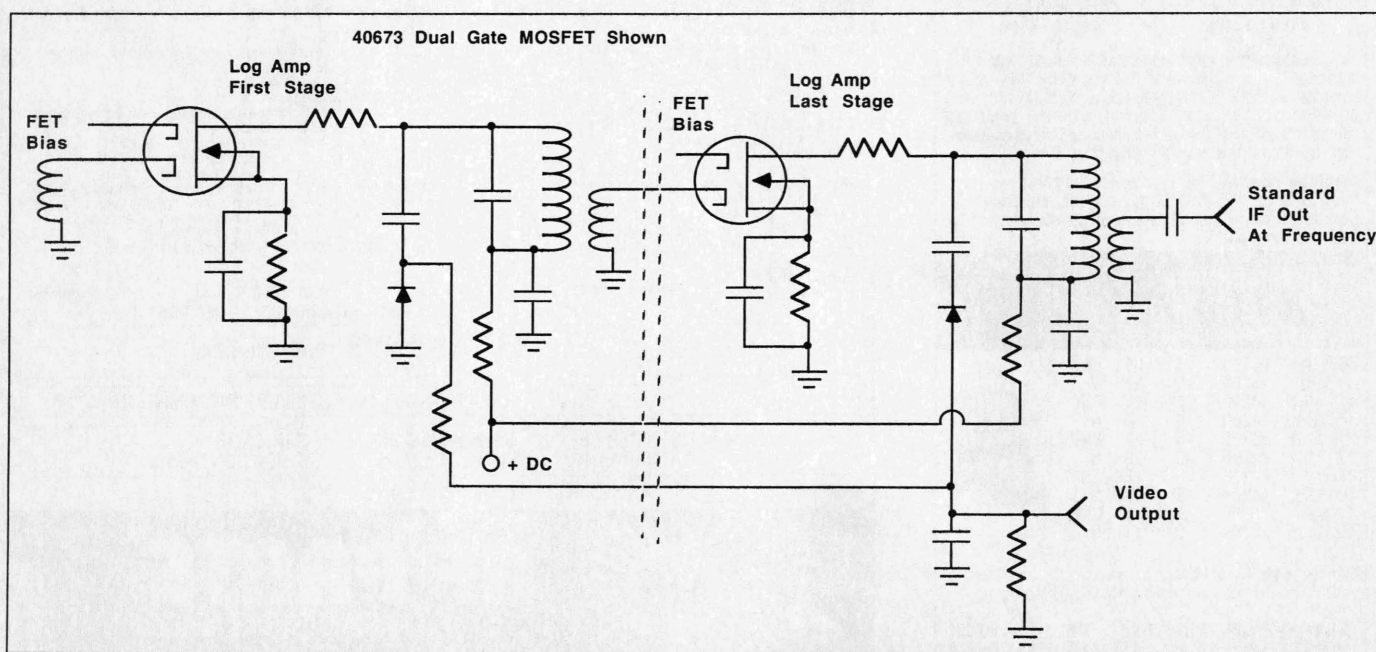


Figure 2. Schematic detail of diode log amplifier showing interconnections. There are usually 6 to 8 stages in cascade.

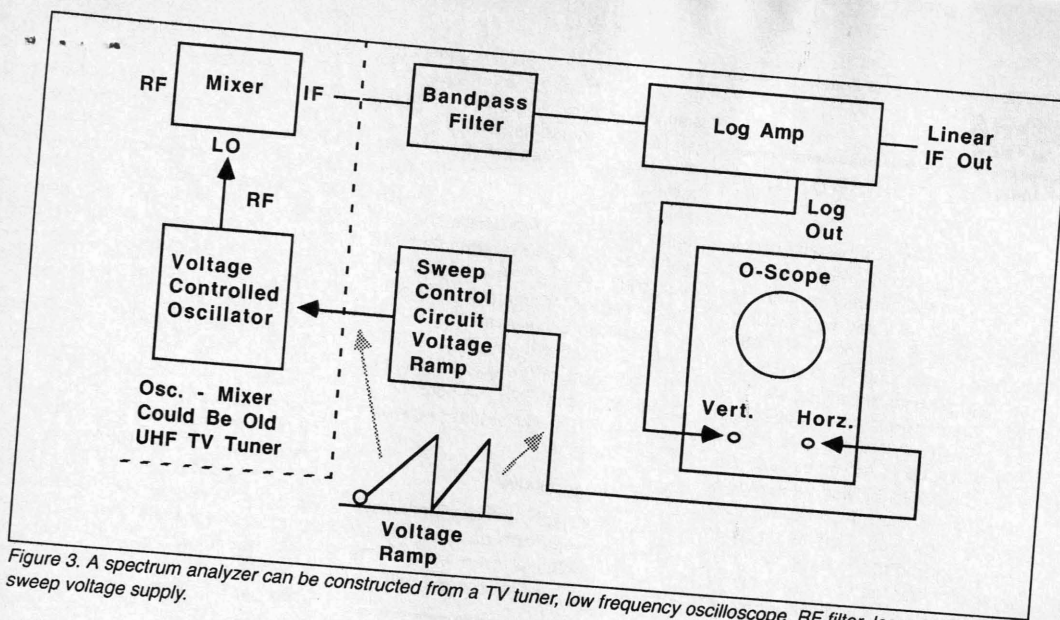


Figure 3. A spectrum analyzer can be constructed from a TV tuner, low frequency oscilloscope, RF filter, log amplifier, and a sweep voltage supply.

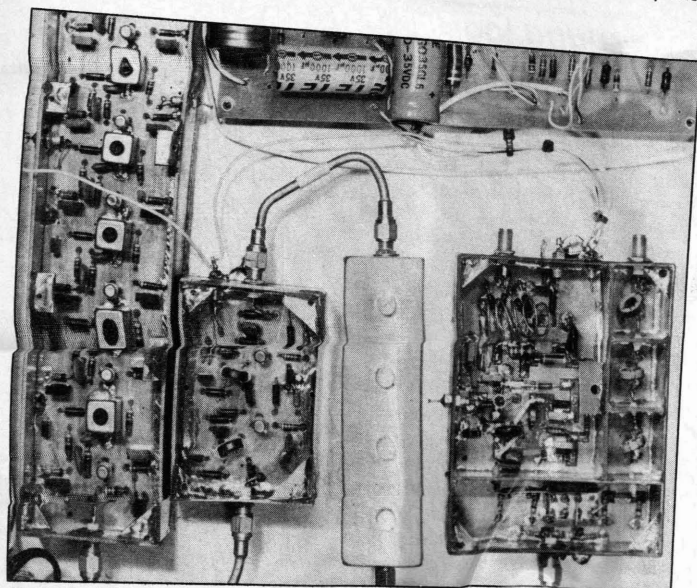


Photo D: Old original home-built spectrum analyzer showing five stages of log amplifier, power sweep circuits, RF preamp and filter, and VCO mixer circuitry.

you will appreciate its dynamic range, abilities, and usefulness in spectrum analyzer applications. I mentioned earlier that with a log amp, a filter, and an RF amp/local oscillator you can make a small unit that is capable of display-

ing data on almost any type of oscilloscope. However, as I stated earlier, a long-persistence type unit would be best, similar to a heart-rate display monitor.

The basic system for a spectrum

analyzer can be easy to set up. Using a TV tuner front-end removes the construction of both the tracking RF amplifier and oscillator requirements. Now only the filter and log amplifier units are required if you can find a medical heart rate monitor (it has the sweep circuits). Hope you have good luck checking out your local swap meets and surplus material sources for equipment.

Mailbox Comments

Sean KB8JNE of Hilliard, Ohio, writes that he wants me to mention in the column that there must be quite a few Columbus, Ohio, area amateurs with 10 GHz equipment in their basement or attic. He is interested in seeing if those who have equipment are willing to dust it off and give it a try again. He says most of the folks around there have played with it (10 GHz) at one time or another, but found the band empty in the area and filed the equipment away. If that's the case, get the equipment out and contact Sean KB8JNE, 3700 Westbrook Drive, Hilliard OH 43026. He would like to organize a local microwave enthusiast group in the Columbus area. Drop him a line. Sean, I hope you get a good re-

sponse to your efforts.

Arthur W1PXL wrote to inquire about an antenna noise bridge for VHF/UHF use and asked about schematic information on the construction of such a unit. Well, there is an antenna noise bridge for 1.8 through 30 MHz in the *ARRL Handbook*, 1989 edition, pages 25-32. The same circuit appears in the *ARRL Antenna Handbook*, 1988 edition, pages 27-15. A similar circuit appears in the *RSGB Handbook*, 4th edition, 1983, pages 11.19-11.20. This unit covers low frequency to 200 MHz and will give good results to 432 MHz. The books from the RSGB, Radio Society of Great Britain, are quite good and describe almost all circuits in great clarity of construction. Most seem to be based on a heavy involvement in home construction. They're good books to have on the bookshelf.

Carl AA4H writes that he recently purchased a multimode VHF transceiver. He found many stations to work on 6 and 2 meters and a fair amount of activity on 70 cm; however the activity level on 222 is very disappointing. He reports making only a few QSOs on 222 MHz and most were in the UHF contest except for a few locals. Why is there such a lack of 222 MHz activity? Is it because of the recent loss of the bottom 2 MHz? Carl wants others to use this band as he feels we will lose further parts of the band if we do not increase our activity in this area. For further information, contact Carl AA4H at 5971 Hwy. 126, Blountville TN 37617.

Well, that's it for this month. Next month I plan to start a construction project modifying an SSB transceiver for microwave use. First I will present a basic 28 MHz platform constructed dead-bug style, showing a simple system. Later we will cover a converter to 2 meters using two of the Hamtronics modules to wrap up the 28 MHz to 2 meter SSB transverter. See you next month. As always, I will be glad to answer questions on this or related topics. Please send an SASE for prompt response—family, contest and workbench time permitting. 73 Chuck WB6IGP.

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CIRCLE 56 ON READER SERVICE CARD

Low-Cost Receiver for Satellite TV

— this modular design uses
readily available technology

Author's Note: This article was originally presented by the author as part of the professional program, "Satellite TV and the Private User," at WESCON/79 in San Francisco CA on September 20, 1979. WESCON, the annual Western Electronics Show and Convention, is a large-scale trade show which features films, exhibits, and professional sessions related to all facets of the electronics industry.

The technological revolution which is making possible the distribution of television programming via satellite has been discussed in both popular and tech-

nical publications and at conferences and seminars.¹ In this article, I will explore the trade-offs involved in designing a wideband, tunable FM video receiver for processing and displaying DOMSAT (Domestic Communications Satellite) signals.

It should be recognized that there are at least as many conflicting receiver design philosophies as there are microwave engineers, and no claim is made that the concepts presented here are necessarily superior to any other approach. Nevertheless, this receiver does provide adequate performance at low cost, and the trade-offs encountered are typical of those with which others have had to deal. It is hoped that documenting this effort will help to dispel some of the

mystique of microwave receiver design.

Signal Characteristics

Unlike the vestigial-sideband AM video standard used for terrestrial TV broadcast, DOMSAT video incorporates a wideband FM format, with audio multiplexed onto a subcarrier prior to modulating the composite. The resulting wideband channel (see Table 1) affords considerable "FM advantage" (signal-to-noise enhancement for a given carrier-to-noise ratio); however, the bandwidth and format tend to complicate the receiver design task.

Were a signal consisting of vestigial-sideband AM video with intercarrier narrowband FM audio available from the satellites, receive processing

Video Carrier	
Channels	24
Adjacent channel spacing	40 MHz
Orthogonal channel spacing	20 MHz
Frequency band	3.7-4.2 MHz
Peak deviation	10.25 MHz
Max. video frequency	4.2 MHz
Pre-emphasis curve	CCIR 405-1
Audio Subcarrier	
Frequency	6.8 MHz
Peak deviation	75 kHz
Max. audio frequency	15 kHz
Pre-emphasis time const.	75 usec
Energy Dispersal	
Waveform	Triangular
Frequency	30 Hz
Peak deviation	750 kHz
Composite	
EIRP	+ 65 dBm
Path loss	- 196 dB
99% power bandwidth	36 MHz
Received spectral density	- 206 dBm/Hz

Table 1. Typical DOMSAT signal characteristics.

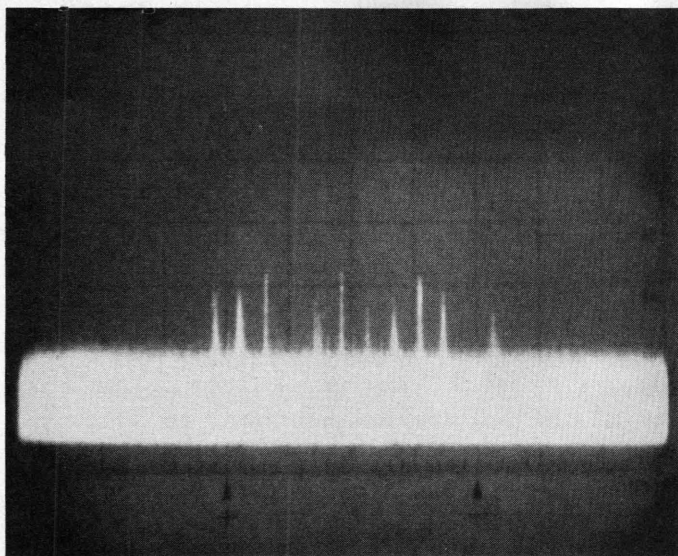


Fig. 1. Spectral display of a 4-GHz DOMSAT downlink recovered on a 4.7-meter antenna and amplified by a GaAs FET Low-Noise Amplifier (LNA). Horizontal deflection is 100 MHz/div, and vertical sensitivity is 10 dB/div. Eleven video carriers, along with their associated FM sidebands, are visible. Note that the fourth channel above the bottom of the band is vacant. Otherwise, channel spacing is 40 MHz and carrier-to-noise ratio appears to be on the order of 10 dB.

would involve merely heterodyning the selected channel in the 4-GHz transmission band against a stable microwave local oscillator (LO), and applying the VHF difference signal directly into the tuner of a conventional TV set. Unfortunately, with DOMSAT signals as they are currently formatted, such a down-conversion process would merely spread unintelligible sidebands across six adjacent TV channels.² Thus, it becomes necessary to design a complete receiver, including heterodyne conversion stages, demodulators, and video and audio processing circuitry, to recover and display satellite TV.

Frequency Agility

It will be noted from Table 1 that the downlink band used by most North American DOMSATs is 500 MHz wide, and that for a given antenna polarization there will be present up to twelve video carriers, spaced 40 MHz apart (see Fig. 1). That these signals

are of extremely low amplitude complicates the design of the Earth station's antenna³ and low noise pre-amplifier,⁴ but we will assume for the moment that an adequate signal-to-noise ratio exists at the input of the receiver to permit signal recovery. The problem at hand, then, is to select a particular 40-MHz wide channel from among 12 such signals in a 500-MHz wide band, while adequately attenuating the adjacent channels.

A Tuned Radio Frequency (TRF) approach, with detection occurring directly at the downlink frequency, would require readily-tunable bandpass filters of high Q (to accommodate the 1% or so channel bandwidth) and skirts steep enough to reject adjacent channels. Tuning requirements rule out both LC and resonant cavity filters, suggesting the use of Yttrium-Iron-Garnett (YIG) sphere resonators for channel selection.

Although YIG filters can readily be bias-tuned, their

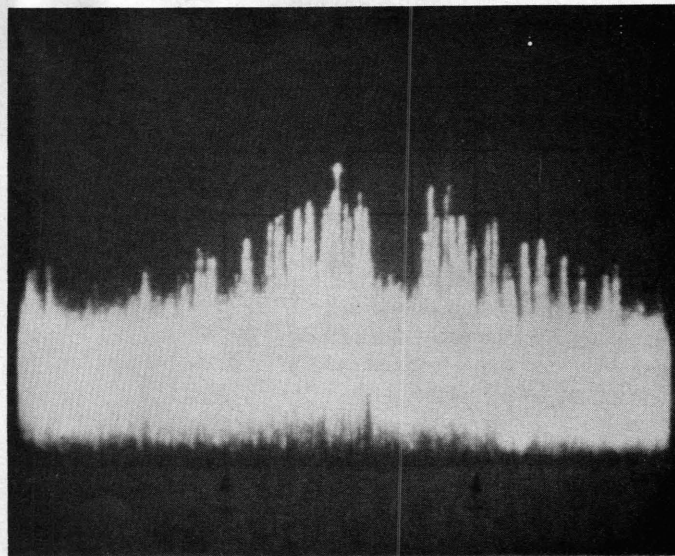


Fig. 2. Spectral display of a single wideband FM video channel after dual-downconversion to a 70-MHz second i-f. Horizontal deflection is 3 MHz/div. and vertical sensitivity is 10 dB/div. This is the composite FM signal from which video and audio are to be demodulated.

cost and the complexity of the required driving circuitry tend to rule them out for private terminal applications. Furthermore, it is far easier to tune a single oscillator than a bank of filters. This suggests heterodyne-downconverting a selected channel into a fixed intermediate frequency, at which demodulation may take place.

I-f Selection

The selection of intermediate frequencies for superheterodyne receivers involves careful attention to the required and realizable mixer bandwidths, image rejection criteria, demodulator circuit capabilities, and tuning constraints. These various considerations tend to be mutually exclusive, but it has been shown⁵ that for narrowband systems, a reasonable compromise is achieved by selecting an intermediate frequency approximately one-tenth the frequency of the incoming signal. Although DOMSAT video hardly qualifies as a narrowband service, we can use the one-tenth rule of thumb to establish a starting point. For a 4-GHz

input signal, this suggests a UHF i-f. However, the various demodulator circuits compatible with wideband FM video (quadrature detector, ratio detector, Foster-Seely discriminator, phase-locked loop and the like) are all most readily realized in the lower portion of the VHF spectrum. An obvious solution is to utilize dual downconversion, with first and second i-fs near 400 and 40 MHz, respectively.

In fact, numerous experimental DOMSAT video terminals have adopted the above frequency scheme, many employing UHF TV tuners for the second down-conversion. The drawbacks to such an approach include the typical UHF tuner's restricted channel bandwidth, relatively high noise figure, and poor local oscillator stability. Nevertheless, when cost is the primary design constraint, these problems can be circumvented.

Not so readily resolved is the input filtering requirement which such a frequency scheme imposes. Assuming low-side first LO injection and top-channel reception, the first conver-

MX-4200 DOUBLE-BALANCED MIXER

Input frequency	3.7–4.2 GHz
LO frequency	2.5–3.0 GHz
Intermediate frequency	1.2 GHz
Isolation	20 dB
Conversion loss	7 dB

LO-3000 VOLTAGE CONTROLLED OSC

Output frequency	2.5–3 GHz
Output power	+7 dBm
Spurious rejection	20 dB
Tuning voltage range	3–10 V dc
Supply potential	+13.5 V dc

RF-1200 AMPLI-FILTER

Center frequency	1.2 GHz
3-dB bandwidth	50 MHz
Gain	15 dB
Noise figure	2 dB
Supply potential	+13.5 V dc

LO-1270 LOCAL OSCILLATOR

Output frequency	1270 MHz
Stability	$\pm 0.001\%$
Power out	+7 dBm
Spurious rejection	40 dB
Supply potential	+13.5 V dc

MA-1200 MIXER-AMPLIFIER

Input frequency	1200 MHz
LO frequency	1270 MHz
Intermediate frequency	70 MHz
Conversion gain	20 dB
3-dB bandwidth	40 MHz
Supply potential	+13.5 V dc
Isolation	20 dB

Table 2. Typical parameters for conversion modules.

sion will generate an image frequency which falls a mere 300 MHz below the bottom edge of the downlink passband. An input filter capable of providing adequate passband flatness and minimal insertion loss over the 3.7- to 4.2-GHz band is unlikely to provide adequate image rejection if a 400-MHz first i-f is utilized. One may wish to raise the first i-f high enough to

separate the image frequency band well away from the downlink passband, thus simplifying input filtering.

In fact, if the Low-Noise Amplifier (LNA) which precedes the receiver utilizes a waveguide input, then an image filter already exists. Rectangular waveguide is a high-pass transmission line. If low-side first LO injection is used and the first i-f is carefully selected, the

LNA's waveguide input will itself reject the image frequency.

Most commercial LNAs utilize an EIA standard WR-229 waveguide input. This guide has a lower $TE_{1,0}$ cutoff frequency near 2.5 GHz. This cutoff frequency is about 1.2 GHz below the bottom edge of the receiver's required passband, so input losses will be minimal. But a first i-f of, say, 1.2 GHz, will place the image frequency as far below cutoff as the input passband is above cutoff. The image thus ends up quite far down the waveguide high-pass filter's skirts, and may effectively be ignored.

True, the fixed 1.2-GHz first i-f requires that the first LO be tunable, but we mentioned earlier that it's far easier to tune a single oscillator for channel selection than a bank of filters. And even at the top of the downlink passband, where the first LO must be tuned up to 3 GHz, the image at 1.8 GHz is sufficiently far below cutoff so that a 12-cm long input waveguide will afford on the order of 60 dB of image rejection.⁶

Another signpost pointing to the selection of 1.2 GHz as a first i-f is realizable amplifier Q. The 3-dB bandwidth of the i-f amplifier string must be greater than or equal to the

20-dB channel bandwidth in order to avoid unduly attenuating significant sidebands. Assuming a channel bandwidth of 40 MHz and an i-f of 1.2 GHz, this dictates an effective first i-f Q of 30. This value is readily realized with microstripline circuitry.

Despite the obvious economic advantages of the modified UHF TV tuner conversion scheme, it was decided to employ a 1.2-GHz first i-f in the Microcomm DOMSAT video receiver. But what of the second i-f—is it similarly constrained by the wide downlink passband? Actually not. With channel selection occurring in the first downconversion, the second i-f need only be wide enough to accommodate a single video channel. Downconverting the 1.2-GHz first i-f to any desired VHF frequency will allow ample second-conversion image rejection with simple i-f filtering while providing adequate bandwidth to pass the 40-MHz composite.

Since the communications industry has long utilized 70 MHz as a standard i-f for microwave links, it was decided to employ a 70-MHz second i-f in the DOMSAT video receiver. This makes it possible to utilize any of the readily available 70-MHz wideband FM i-f strips to demodulate the video information.

Gain Distribution

Gain partitioning for the DOMSAT video receiver depends upon the available power from the satellite, the threshold sensitivity of the demodulator circuitry selected, and the gain of the receive antenna utilized. It has been shown that for the illumination contours typical of most North American DOMSATs, an optimum private-terminal antenna will exhibit on the order of +41-dBi gain.⁷

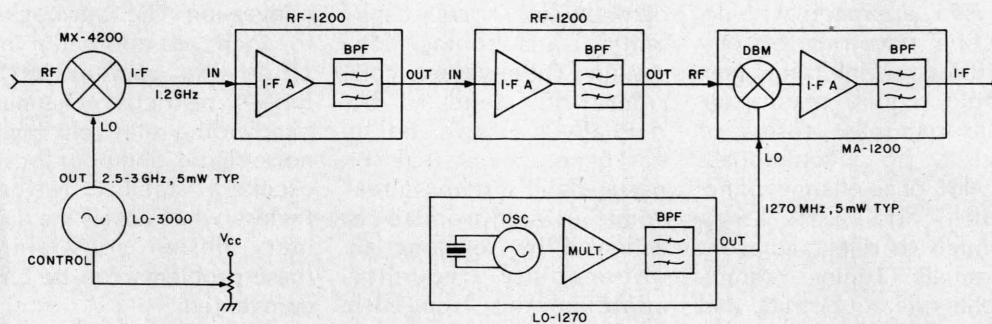


Fig. 3. Block diagram of the heterodyne downconversion portion of the DOMSAT video receiver. Shown at left is the 3.7- to 4.2-GHz input terminal from the LNA and feedline. The 70-MHz i-f output at right feeds the baseband demodulator and processing circuitry (see Fig. 7). The potentiometer shown at the lower left represents the resistive voltage divider which tunes the first LO for channel selection. Gain partitioning of the various blocks is discussed in the text.

Given the EIRP and path loss numbers listed in Table 1, it appears that the signal level available to the LNA will be on the order of -90 dBm.

The input threshold for a typical phase-locked loop (PLL) integrated circuit operating as an FM demodulator at 70 MHz is on the order of -20 dBm. This suggests that between the antenna and the demodulator, roughly 70 dB of conversion gain is required.

There are three sources of gain available between the antenna and the PLL. These include the LNA and first and second i-f amplifiers. There are, similarly, three sources of loss in the system: the insertion loss of the transmission line which connects the LNA to the receiver, and the conversion loss of the first and second mixers. For a typical home installation, the feedline insertion loss may be on the order of 6 dB, and if double-balanced diode mixers are used for the two frequency conversions, it is safe to assume that the conversion loss of each will be on the order of 7 dB. This suggests that the overall gain of the LNA, first, and second i-f amplifiers will need to total 90 dB for adequate DOM-

SAT video reception.

In the interest of maximizing system stability and dynamic range, it is desirable to distribute the required 90 dB of gain uniformly between the rf and two i-f frequencies. A 30-dB gain LNA is clearly feasible at 4 GHz and would require three stages of GaAs FET amplification. This amount of LNA gain is sufficient to adequately mask the noise temperature contribution of the feedline and receiver, allowing the low-noise temperature of the FETs to predominate.⁴ Similarly, it is practical to achieve the desired 30 dB of 1.2-GHz gain by cascading two stages of ion-implanted silicon bipolar transistor amplification. At 70 MHz, the required gain is readily available from thin-film wideband gain blocks produced by a number of different vendors.

A block diagram for the dual downconversion portion of the DOMSAT video receiver, partitioned in accordance with the foregoing discussion, is shown in Fig. 3.

Construction of Conversion Circuitry

During the initial system-development phase of any

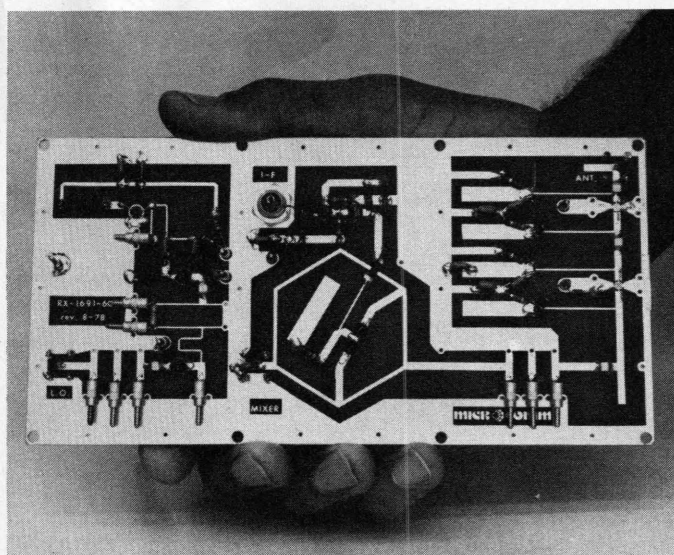


Fig. 4. Typical microstripline circuit module used for satellite video downconversion. Each module in the receive system is mounted in its own shielded enclosure, and all are interconnected via coaxial cable as discussed in the text.

new product, it is common practice to build a number of different amplifier, mixer, filter, and oscillator circuits, each connectorized for coaxial input and output and with each circuit separately boxed and shielded. A modular developmental system provides the engineer with the flexibility of changing one or more circuits without having to disrupt the rest of the system. Microcomm's earlier efforts at modular receiver development have been documented pre-

viously.⁸

But modularization has advantages for a production system as well. If every function represented by a block in Fig. 3 is implemented in a separate, shielded module, then isolation between stages is maximized and the crosstalk and stability problems associated with stray rf coupling can be eliminated entirely. Further advantages are realized in the area of maintainability. Should a receiver fail, fault isolation by module sub-

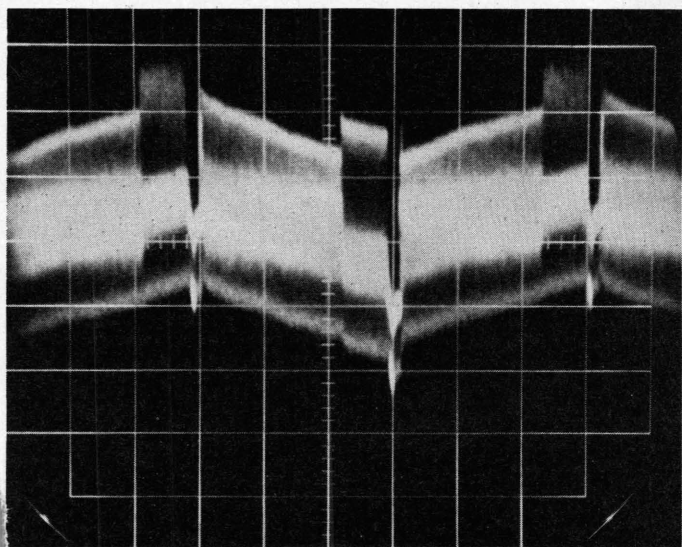


Fig. 5. Baseband output of the PLL demodulator. Presence of the energy dispersal waveform on the video composite is evident.

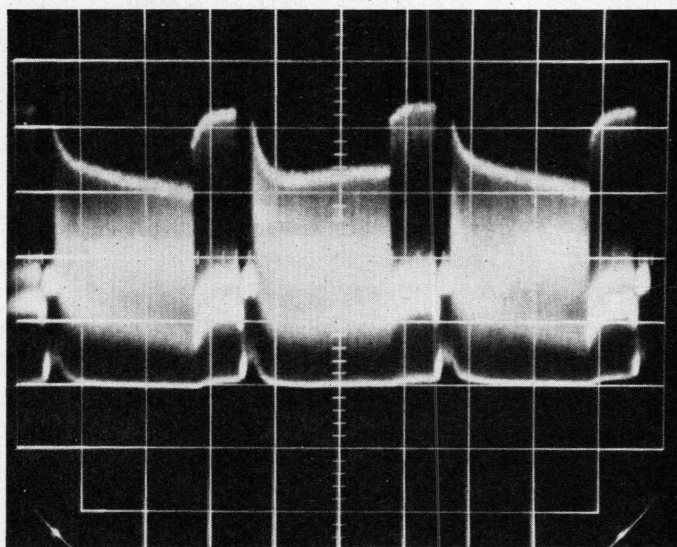


Fig. 6. The dc restorer is simply a diode clamp circuit which removes from the video waveform any vestiges of the energy dispersal waveform seen in Fig. 5.

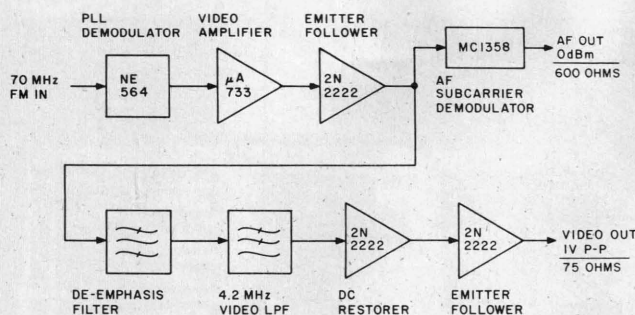


Fig. 7. Block diagram of the complete baseband processing portion of the DOMSAT video receiver. The above circuitry is driven by the output of the downconversion module set (see Fig. 3), and provides standard 1-volt video and 0-dBm audio outputs to an external modulator, studio monitor, or video tape deck.

stitution becomes a viable troubleshooting technique. And, of course, a modular system maximizes user flexibility by allowing customers to assemble from standard modules a custom system designed to meet their precise needs.

The specifications of the modules developed to implement Fig. 3 appear here as Table 2. Each of these modules employs microstripline construction, as shown in Fig. 4, to minimize component count and assure duplicability.⁹

Baseband Processing

Before the wideband FM composite shown in Fig. 2

can be displayed, several processing steps are necessary. The 70-MHz i-f signal will, of course, be demodulated first, and this may be accomplished readily by using a monolithic PLL in a standard circuit.¹⁰ The output waveform from the PLL will contain both the video waveform and the modulated audio subcarrier, but superimposed on these will be found the 30-Hz triangular energy dispersal waveform added to all DOMSAT downlink signals as an interference reduction technique. This waveform is evident in the oscilloscope display in Fig. 5.

Prior to attempting to re-

move the energy dispersal waveform, it is desirable to amplify the rather feeble video level available from the PLL demodulator, and this may be accomplished using a single monolithic TV video-amplifier IC. Next, an emitter-follower permits splitting off the 6.8-MHz audio subcarrier for demodulation in a standard TV sound i-f microcircuit, whose associated circuitry is modified slightly for compatibility with the higher carrier frequency and peak deviation used on satellite audio.

After passing through a de-emphasis filter and passive video low-pass filter, the video waveform may finally be applied to a diode clamp circuit which will remove the energy dispersal waveform (see Fig. 6). An emitter-follower then establishes the desired 75-Ohm video output impedance to drive recording or display circuitry, as required.

A block diagram for a complete baseband processing subsystem is shown in Fig. 7. This circuit can be constructed on a single printed circuit board and incorporated into a complete DOMSAT video receiver by simply interfacing it to the downconversion circuitry shown in Fig. 3.

Display Options

An ideal DOMSAT video receiver for the home Earth-station market would provide an intercarrier audio, vestigial sideband rf output for direct interface to the user's VHF TV receiver. Such an rf output may readily be realized by using any of the available video modulator microcircuits developed for the TV game and home computer industries. In fact, the video and audio levels available from the baseband unit shown here are entirely compatible with such modulators.

Unfortunately, incorp-

orating an rf modulator in a commercial DOMSAT video receiver would subject the entire receiver to FCC type acceptance in the United States. As more than one home computer manufacturer has discovered, the type-acceptance procedure is burdensome in the extreme, with bureaucratic delays often precluding a timely market entry. In addition, the resolution and clarity of most of the available low-cost video modulators leave quite a bit to be desired, and rf modulation would tend to degrade overall video quality noticeably.

A possible solution would be to provide the user with simply a video and audio output from the DOMSAT receiver and allow him to display the receiver's output on a studio-quality TV monitor. However, few videophiles possess such a monitor, and the cost is prohibitive.

Fortunately, most videophiles do possess a videotape recorder (in fact, the owner of a home satellite Earth station would most likely find it impossible to function without one!) and the average video recorder contains an extremely high quality rf modulator. Allowing the user to interface his DOMSAT receiver to the TV via a video recorder provides an ideal solution to the type-acceptance dilemma. And those users who have no recorder are, of course, free to add an external rf modulator, any number of which are available in kit or assembled forms.

Equipment Availability

Once priced in the tens of thousands of dollars, DOMSAT video receivers are now being brought within the reach of the American consumer. The conversion modules shown in Fig. 3, for example, have, since late 1978, been available to the ex-



Fig. 8. The video output of the baseband unit, displayed on a TV set with the aid of the rf modulator in a video cassette recorder. Most DOMSAT viewers find the clarity and resolution of satellite video clearly superior to network video which is distributed by multiple terrestrial microwave hops.

perimeter at \$1000 complete, and to the OEM at significantly lower prices in production quantities.

It is expected that by the end of 1979, a complete, fully tunable DOMSAT video receiver, utilizing the above conversion modules and a baseband unit such as that blocked out in Fig. 7, will be available to the consumer for under \$2000. This receiver will be fully packaged and assembled, including power supply, tuner, control circuitry, and inter-connecting cables. Such a receiver promises to make possible for the first time a complete, consumer-grade Earth station, including antenna and LNA, for under \$4000. We can only begin to guess at the price breakthroughs which may follow! ■

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SKIP CAMPISI

By now, everyone has heard of "gravity waves." But why have they never been detected? The common answer from classical physics is that gravity waves are very "subtle and weak." To me, that answer appears to be a very subtle and weak explanation! To discuss this topic in an intelligent manner, we must first examine some facets of Relativity and Particle Theory, which both tend to be somewhat hazy in the areas concerning gravitation.

Take, for example, Einstein's Principle of Equivalence, on which his Relativity Theory is heavily (if not entirely) based. Here he has stated that there is absolutely no way for an observer in a fully enclosed environment to distinguish whether he is "at rest" in a gravitational field, or is being "accelerated" by a uniform force pushing (or pulling) his enclosure. Picture yourself in a rocket ship at rest on the Earth, and then experiencing a one-gravity rocket acceleration far away from any local gravitational fields.

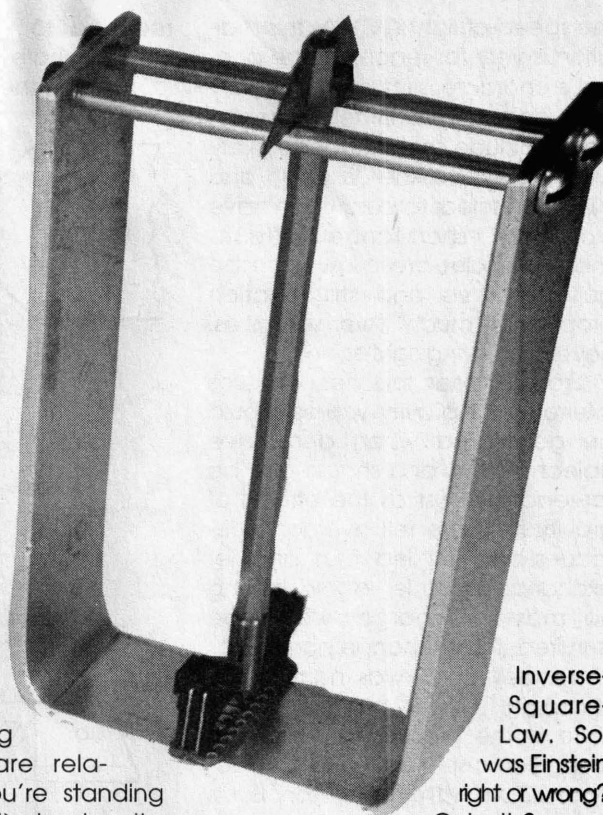
At first glance, you'd agree with Einstein—the one-gravity rocket acceleration would feel exactly the same as sitting at rest in Earth's gravitational field. However, that's only because your senses are too limited to note any difference. What? You don't believe me? You've forgotten all about the Inverse-Square-Law!

The force of gravity falls off and weakens as the distance to the

mass creating the gravitational field increases, adhering to an inverse-square relationship. Thus, if you're standing upright (in the rocket) at rest on the Earth, your feet are experiencing a slightly stronger pull from gravity than your head, agreed? What happens to you when the rocket is accelerating outside of a gravitational field?

Naturally, your head feels the exact same pull as your feet. As you can readily see, there is a difference between gravity and acceleration. But how do you measure this minuscule little difference? Fasten your seatbelts—here comes a very subtle contradiction in Relativity: Einstein himself gave us the means to make the measurement! In explaining his theories, he loved to use measuring rods and clocks. According to him, a clock placed within a strong gravitational field would run slower than one placed in a weaker gravitational field.

This aspect of Relativity was confirmed in 1960 at Harvard University using the Mossbauer Effect, which utilizes atomic nuclei as extremely accurate "clocks." The "clock" on the bottom floor of their lab ran slower than the one on the upper floor. Obviously, this result won't occur in an accelerating rocket, where the force is uniformly applied and not subject to the



Inverse-Square-Law. So, was Einstein right or wrong? Or both?

Technically speaking, he was correct, as numerous proofs of his Special and General theories have upheld all of his assumptions to date. His Principle of Equivalence indeed appears to hold true, even though he himself has shown us that there is a method of distinguishing between uniform acceleration and gravitational force. What I'm attempting to point out here is that at least one stone has been left unturned. Perhaps we'll find others in our quest to understand gravitation.

Another peculiar aspect of Relativity is the so-called limiting velocity: A particle with mass can only attain speed-of-light velocity by acquiring infinite mass from infinite energy input, which are both impossibilities. This indeed appears to be true according to particle-accelerator experiments. But what about particles without mass? This is where we depart from Relativity and enter the equally strange world of Particle Theory.

Here we use the photon as an example. The photon is (according to current theory) the "exchange particle" for electromagnetic radiation. It has no charge or mass, and a "spin" of 1. It travels at one speed:

the speed-of-light (1C). As this radiation is very far-reaching, the photon's characteristics are necessary for propagation. (Other examples could include the exchange particles related to the "strong" and "weak" nuclear forces, which have extremely short-range effects. These particles are allowed to be quite massive and still function properly at much lower velocities, never achieving light-speed.)

From these theories, it was inferred that gravity waves should be generated when a massive object moves, and should also be far-reaching, just as the effects of gravity itself are felt over long distances. This implied that another exchange particle, again having no mass or charge, would be required. This exchange particle, as yet undetected, was named the "graviton."

To fit the theories, the graviton was given a spin of 2, and a velocity of 1C, as with the photon. But is that correct? What if the spin is actually 1 and the velocity equal to 2C? I hear you all grumbling—nothing can exceed light-speed, can it? Nothing with mass, that is. Let's try a little experiment (see Fig. 1) that appears to thwart Relativity a bit.

For this experiment, we'll need a fixed reference frame, which is actually simpler than it sounds. All that we need to do is attach two HeNe lasers onto a length of board so that they face each other, with their beams aligned parallel, traveling in opposite directions to each other. The beams are aligned so that they can't enter the emission aperture of the opposing laser, which could cause unstable lasing. A target is placed next to each aperture to display the opposing laser's beam.

If you turn both lasers on, you'll notice the beams' images displayed on their respective targets. If you now turn off only one laser, you'll notice that the remaining beam display has not changed in intensity or color. This result has only two possible explanations:

- 1) The beams passing in opposite directions interacted with each other in a "Relativistic" manner. The velocity of each beam was

reduced to 0.5C, for a relative velocity shared by both beams of 1C, while the laser frequencies

were doubled (wavelengths halved) to maintain the original color and intensity. (Difficult to

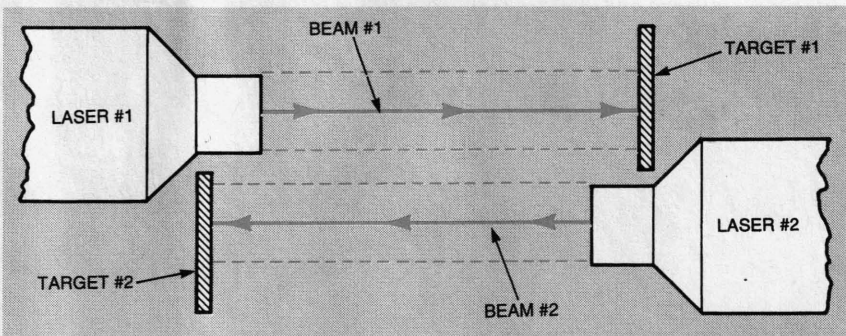


Fig. 1. This simple experiment in Relativity yields two possible conclusions that have profound implications.

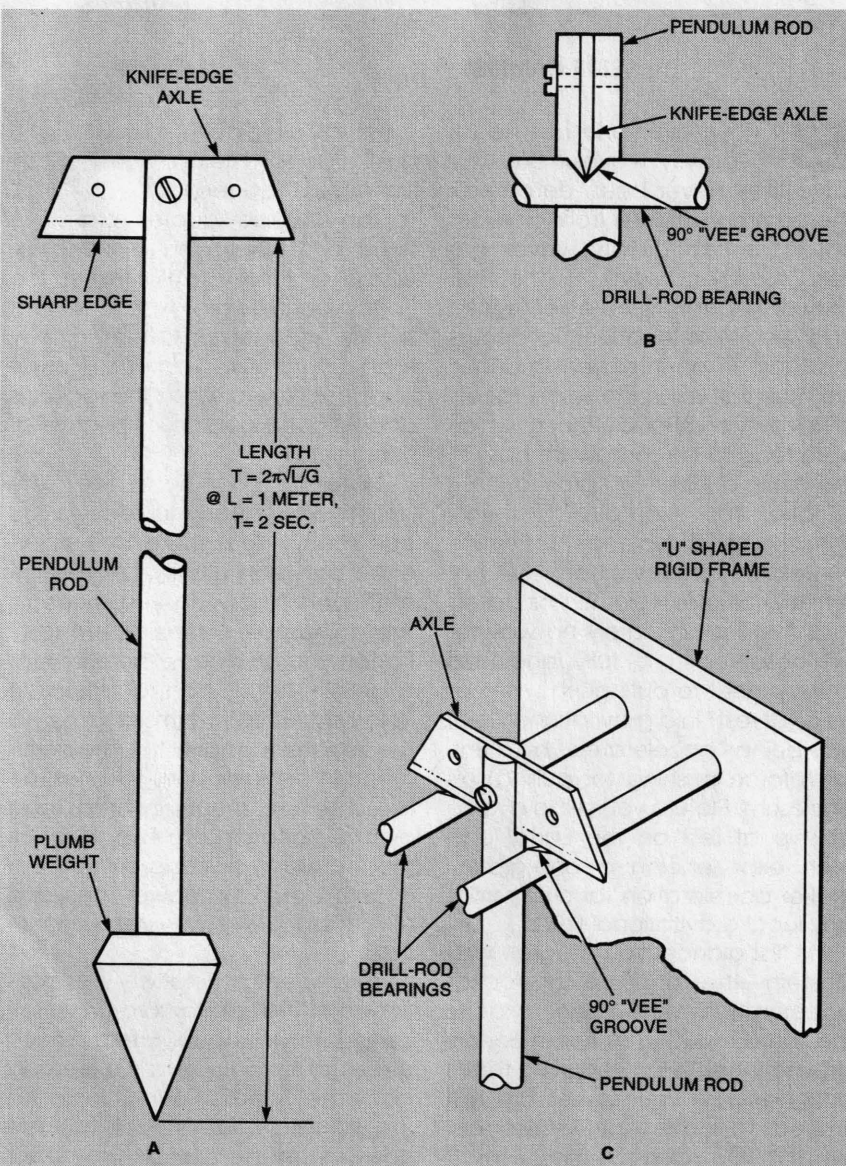


Fig. 2. Here is one possible method (using commonly available materials) of constructing a super-sensitive pendulum movement. The key concerns in any pendulum design are low friction and rigidity of components. The pendulum itself is shown in A, a side-view of the axle and bearing details is shown in B, while the sectional view in C shows the basic assembly.

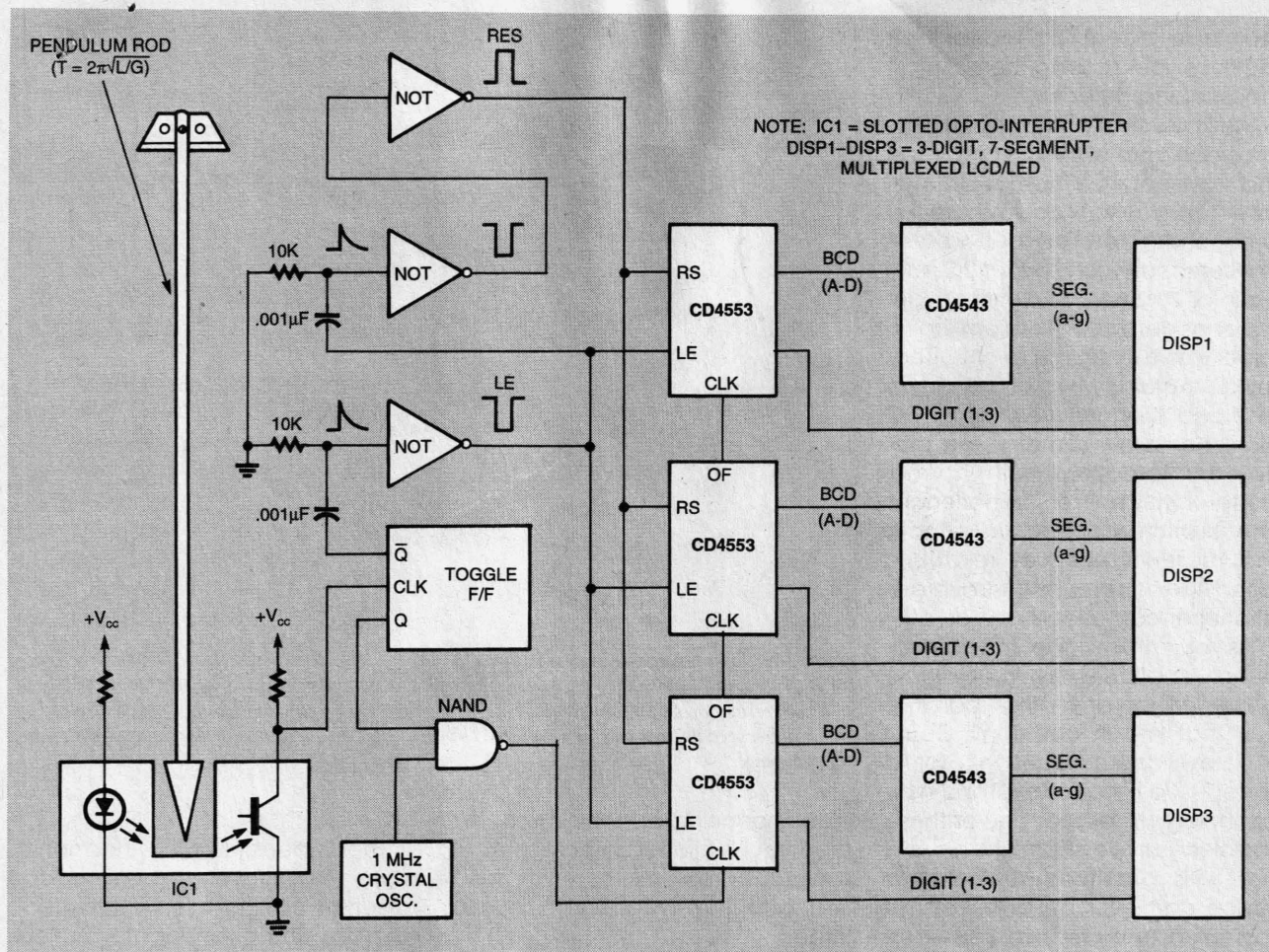


Fig. 3. As shown in this block diagram, the tip of the pendulum interrupting the light beam of IC1 provides a "gate" signal that triggers the display, which can be a standard frequency counter set up to read time period. Optionally, you could build your own high-resolution display as illustrated here or use a single-chip version such as the ICM7226A 8-digit counter.

believe, no?)

2) The beams passing in opposite directions had no interaction with each other. This leaves us with the assumption that a photon from one beam shares a relative velocity of $2C$ with a photon from the opposing beam and implies that certain particles are allowed to travel in excess of light-speed without violating any limiting laws. (Any better?)

Both explanations yield profound results: (1) appears to tell us that electromagnetic radiation is always traveling at changing velocities to compensate for the ever-present interaction of photons, reflected or otherwise, and (2) apparently tells us that light-speed can be exceeded by particles without mass. Explanation (1) looks like great fodder for another article,

so we'll leave that alone for now. For the purposes of this article, explanation (2) will suffice.

We know a bit about photons, as they readily interact with normal matter. They don't appear to accelerate or decelerate, but are traveling at light-speed when emitted. One other particle without mass or charge is known to travel at light-speed: the neutrino. This particle was originally "designed" to maintain the Conservation-of-Momentum Law as related to other sub-atomic particles. It has a spin of $1/2$ and is necessarily ejected in certain nuclear reactions to maintain this law, among others.

The neutrino doesn't interact with matter to any great extent, making it undetectable until recently. Some had even speculated that it actually did have a very slight mass, which would not allow it

to reach light speed. Fortunately, Supernova 1987A, located tens-of-thousands of light-years away in the Magellanic Clouds, showed us the truth: neutrinos from this vast explosion arrived at our planet within the same time-frame as the photons from its visible light. If the neutrinos had any mass, we'd still be waiting for them to arrive, due to the enormous distance they would have had to travel at sub-light velocities!

"Okay, so what's all this have to do with gravity waves?" you might now ask, and rightly so.

As I previously stated, the exchange particle theorized for gravity is the graviton. A moving massive object should produce gravity waves using gravitons, just as electromagnetic waves use photons as the exchange particles. Supposedly, gravitons interact to an

even lesser extent with matter than neutrinos, thus making them almost impossible to detect.

We know that photons and neutrinos don't have a mass or charge and travel at 1C. That would indicate that gravitons could share the same characteristics, but what if gravitons really do travel at 2C, with a spin of 1? (Energy-wise, this might be considered equivalent to a velocity of 1C with a spin of 2.) That could explain why we've never detected them: a velocity of 2C might not allow some of the predicted interactions with normal matter (which itself experiences time dilation, contraction effects, and mass increase at relativistic speeds) except for the usual gravitational effects already observed.

As you can imagine, this has profound implications in Particle Theory. Might any other particles exist that are in question? Could be. How about tachyons, for a starter? We'll need a brand-new technology to detect any of these particles. But we still might have a shot with gravitons, assuming a device can ever be built. Let me pass on a few ideas to you—perhaps one of our readers would care to make the attempt at constructing one of these—the successfully completed project would be Nobel material, without a doubt!

One of these ideas is based on the assumption that an incoming gravity wave will interact with our local gravitational field, either adding to it or subtracting from it. Another idea, which is currently popular, assumes that the gravitons interact in a direct manner with matter itself, changing the dimensions slightly (for example) of a solid, material object. The last idea assumes that photon emission wavelength varies with gravitational field strength. Could any of these ideas be sensitive enough to detect gravity waves? Only some R & D will tell!

The major problems with all of these ideas involve the apparent size of the wavefront encountered, and the wavelength of the "radiation." The wavefront, which may possibly be compared to "ocean" waves in a loose sense, is huge—

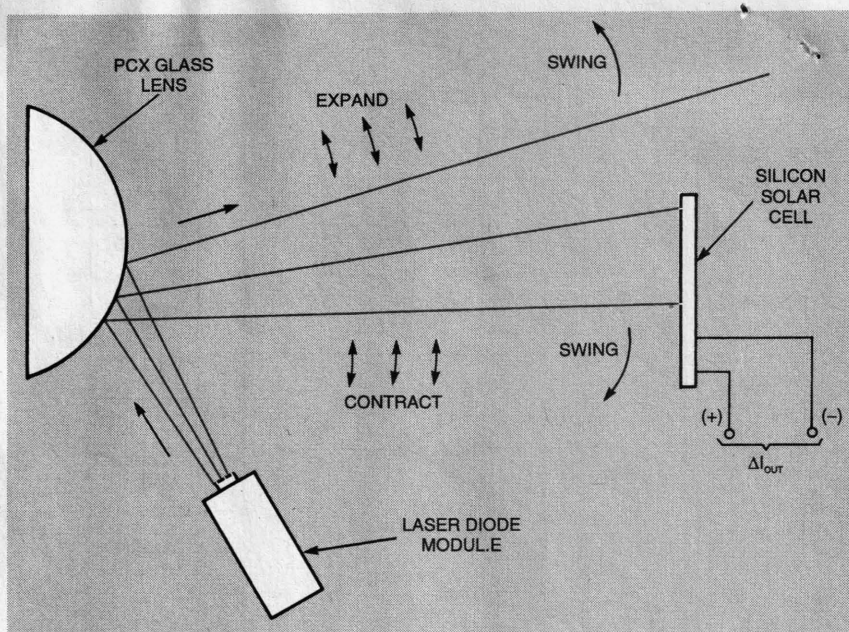


Fig. 4. The "laser bounce" system is perhaps the easiest detector to assemble, requiring only a simple glass lens, laser diode module, and a solar cell. The output current variations can be monitored on a high resolution DMM. Mount all of the components on a rigid frame, with as much distance as practical between the lens and the solar cell. The entire unit should be sitting in a sand box for vibration damping.

globe spanning at a minimum. This creates directional and other measurement problems, as you'll soon find out. The wavelength is also large.

Consider a rapidly rotating "pulsar," which is probably the highest frequency gravity wave generator we might find. If it rotates in one millisecond (note that none have been found this fast as yet), the wavelength would be 186 miles, assuming a propagation velocity of 1C. At a propagation of 2C, the wavelength doubles! We can safely assume that most wavelengths we'll be trying to detect are much longer than 186 miles. So, here we go....

Super-Sensitive Pendulum Method.

Galileo originally determined that the simple harmonic motion of a pendulum resulted in a time period that depended only on the length of the pendulum and the acceleration of gravity. The mass of the pendulum only aids in overcoming friction. If you divide the length by the acceleration of gravity, extract the square-root of the result, and multiply it by 6.28 ($2 \times \pi$), you'll obtain the period of the pendulum. This holds true over short distance swings of about 25% of the pendu-

lum's length.

The pendulum's harmonic motion has proven to be so accurate that oil and mineral prospectors use them to locate buried deposits, which alter the strength of the local gravitational field. Assuming that incoming gravity waves alter local gravity in a similar fashion, pendulum-time-period variations present one possible method for detecting them.

The basic setup is actually quite simple, especially if you already have a frequency counter with time-period measurement capability. The main task then becomes only a matter of assembling a suitable pendulum movement. A suggested approach using common materials is shown in Fig. 2. If necessary, you could construct a high-resolution display (see the block diagram on Fig. 3) using standard logic chips of your choice.

The pendulum itself is fabricated from a rigid length of metal rod or tubing, with a short slot cut down the center of its axis on one end. A "knife-edge" axle (made from a standard "utility" knife blade) is centered in the slot with its edge fixed at 90° to the pendulum rod. Two parallel "bearings" made from drill rod are supported horizontally

and exactly level on a rigid frame, spaced just far enough apart to clear the pendulum. The pendulum swings between the bearings on its knife edge, which pivots in adjacent 90° "Vee" grooves cut into the bearings as "keepers." A massive weight, such as a carpenter's plumb, is installed on the free end of the pendulum rod.

A standard slotted opto-interrupter is located on the bottom section of the support frame (at the center of the pendulum's swing) so that the tip of the plumb can pass through the slot to break the light beam. The pendulum length should be selected to have a period of at least a couple of seconds, and the whole mechanism must sit in a sand box for vibration damping.

The opto-interrupter supplies the gate pulse for triggering the time period display. If you build your own display, you can use a 1-MHz crystal oscillator module along with 3-digit BCD counters driving three BCD-to-7-segment decoders. These in turn could drive three separate, 3-digit multiplexed LED (or LCD) displays, which would display the period in

microseconds. (The ICM7226A counter is a good, single-chip alternative.) You may process the outputs with your PC, utilizing a plotter to provide hard copy of time-period variations.

Needless to say, that method is not useful for higher frequency gravity waves, due to the long period required for the pendulum to give high resolution. Very long-term changes, such as the daily influences of the Sun and Moon, can be easily factored out, especially if you're using a PC. To start the measuring process, gently set the pendulum in motion, giving it a few minutes or so to stabilize before attempting any measurements.

Laser Bounce Method. One school of thought suggests that gravitons may interact sufficiently with matter to actually cause dimensional changes (albeit very small changes) in solid objects. Looking at the diagram illustrated on Fig. 4, you'll find one possible method of detecting these changes.

Here we bounce an uncollimat-

ed laser beam off of a convex reflector, with the reflected beam directed to a large aperture photodetector. To take advantage of "angular magnification," the detector is mounted as far away as practical from the reflector. That allows for a large "moment arm" that functions just as a lever would in a mechanical system due to the convex shape of the reflector.

If the theory holds true, the dimensions of the reflector should change in step with an incoming gravity wave, causing the beam to move across the detector. It can also be inferred that the beam's angular diameter would vary due to the reflector's changes, providing additional amplification for the system. If we locate the detector in the outer fringes of the diverging laser beam, any beam movement would cause the detector output to vary as the gravity waves impinge on the reflector.

The reflector itself should have a precision-polished optical surface. A simple, uncoated plano-convex (flat on one side, convex on the other) glass lens of short focal length should perform well. Even though bare glass only reflects about 5% of the incident light, that is more than sufficient for our purposes. A standard silicon "solar cell" (shielded from ambient light) should make a reasonable detector, using a visible or infrared laser diode as a light source. Once again, the entire system must sit in a sand box for vibration damping, with all components mounted on a rigid frame.

Unlike the mechanical system discussed earlier, this system can handle higher-frequency waves, which might be a great advantage when considering what types of gravity-wave generators might produce the strongest radiation—the fastest moving objects! The output current from the detector can be processed by a PC and plotted for hard copy if desired or displayed on a high-resolution DMM.

Optical Heterodyning Method. Einstein's Theory of Relativity also predicted *gravitational redshift*. All atoms emit photons at specific wavelengths related to the vibra-

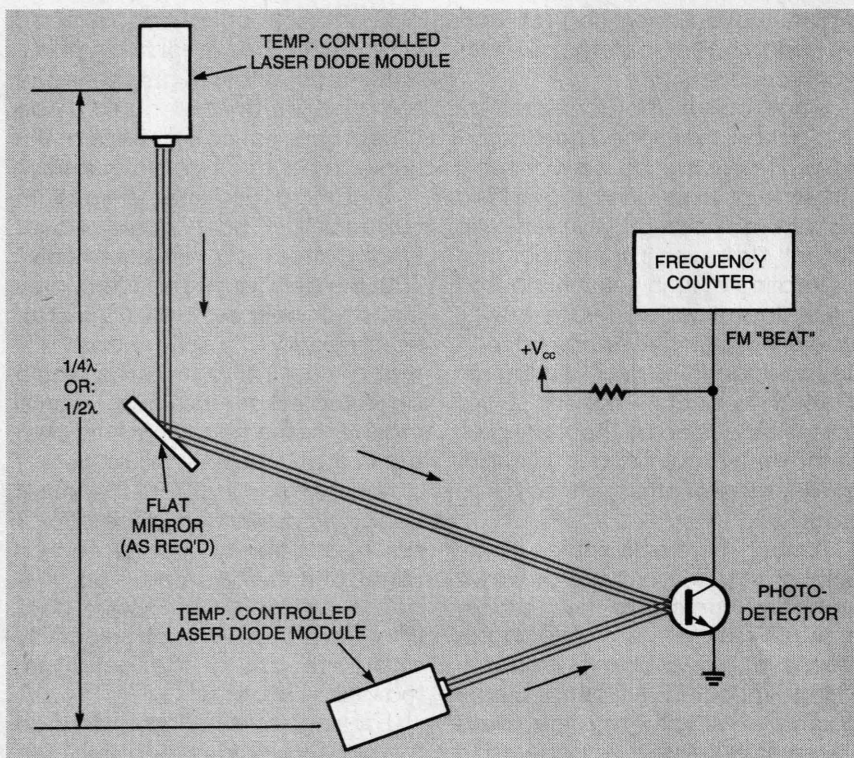


Fig. 5. A promising method of detecting gravity waves uses this unusual type of interferometer. Two temperature-controlled laser diodes apply their beams to a photodetector "mixer" where they are heterodyned. The incoming gravity wave might then frequency modulate this "beat" frequency, as each of its "crests" and "troughs" sequentially pass by the two lasers.

tional rate of their orbital electrons; thus, they can be considered as "clocks." Recalling the 1960 Harvard experiment previously discussed, you'll see why a photon emitted in a strong gravitational field would have a longer wavelength than one emitted from a similar atom in a weaker field. Hence the term gravitational redshift. This has been observed in photons from our Sun.

If we assume that gravity waves affect these "atomic clocks" in a similar manner, we can construct a device such as the one illustrated in Fig. 5, which theoretically could detect the resulting redshift changes. This device is an unusual type of interferometer that uses two light sources, rather than bending the beam back from a single source onto a target.

The system operation is quite simple: One laser diode directly illuminates a photo-detector in close proximity. Another laser diode is located at an optimum distance from the detector, at the point where the greatest amplitude difference would occur (between the two lasers) on the incoming gravity wave. This would happen at one-half of the wavelength, with one-quarter of the wavelength being a more reasonable "antenna" length. As you can appreciate, this method is highly directional and tunable to the desired gravity-wave frequency and wavelength.

With both beams illuminating the detector, which can now be called a "mixer," optical heterodyning results. The mixer's output is the difference frequency between both lasers. If we assume that both lasers' emissions are undergoing redshift changes via the incoming gravity wave, we should be able to see a modulation in this "beat" frequency in step with the gravity wave's frequency as each "crest" and "trough" passes the lasers sequentially.

Selecting a reasonably low beat frequency might allow you to measure that frequency-modulation product. A standard frequency counter can be used as the display, with your PC processing and plotting the results. You might now be wondering how to select two

laser diodes that are close enough in wavelength to have a beat frequency low enough to use conventional circuitry. The answer is that you don't really have to!

It is well known that a laser diode's efficiency goes up as its temperature goes down, and also, its wavelength decreases as its temperature drops. Thus, we have a practical method of "tuning" each laser's wavelength or frequency. By keeping the remote laser's temperature constant, while varying the other laser's temperature, we can actually select an appropriate beat frequency for display. As with any interferometer, vibration-free component mounting is an absolute must!

Detection Problems. At first glance, those three ideas appear to be workable methods for detecting gravity waves, assuming that the waves actually do exist in the first place! We've already pointed out the three major problems concerning detection: the huge wavefront, extremely large wavelengths, and the possibility of faster-than-light propagation velocities. The types of gravity-wave-generating objects we might expect to detect are also worth considering.

What about that one-millisecond pulsar I mentioned previously? It would certainly be convenient if that type of "neutron star" provided us with a 1-kHz gravity wave to detect! However, just the fact that it's rotating at high velocity doesn't mean that it will generate gravity waves. It would necessarily have to be lop-sided in its mass distribution for that to occur. That this is the case is suggested by the X-ray pulses that they radiate, but that fact by no means offers an absolute proof.

A much more likely source of reliable gravity wave generators will probably turn out to be variable stars. These fall into two basic groups: eclipsing binaries and pulsating variables. Eclipsing binary stars are rapidly rotating pairs whose orbital planes lie edge-on to us. Their periods are measured in days, which would make gravity-wave detection a chore. On the other hand, pulsating variable stars

FOR MORE INFORMATION

Both Relativity and Particle Theory are fascinating subjects, with many books having been published concerning them and their relationship with gravitation. If you wish to learn more about these subjects, in the opinion of the author, the following texts are among the best available and can be located through most book-sellers:

Relativity, the Special and the General Theory, by Albert Einstein,

Atom : Journey Across the Subatomic Cosmos, by Isaac Asimov

Conceptual Physics, by Paul G. Hewitt

have periods measured in hours, with the fastest in minutes—much more reasonable! As you can see, a time-period display (as illustrated in Fig. 2) might be a better choice than a frequency display.

I'm sure that you'll now be wondering how a single star that only varies in brightness (magnitude) can produce gravity waves of any strength without moving to any extent. The fact of the matter is that they do move! According to current well-established theory, these stars actually contract and expand in sync with their magnitude pulsations. *Ergo*, they might generate something that could be detectable. So, how does all of this relate to our three detector ideas?

Returning to the pendulum method, just how would wavelength affect our measurements? With the pendulum and the display circuitry in near proximity (as dictated physically) it's quite a possibility that a local 1-MHz oscillator would be affected in the exact same manner as the pendulum time period. In that case, we would detect absolutely nothing! On the other hand, the oscillator may be affected in a different manner, which might effectively amplify the time period variations. This system does offer some directional selectivity along the axis of the pendulum rod—if it works at all.

Looking at the laser bounce method, it would appear that just by adding more convex mirrors to the light path, we could amplify even the smallest dimensional

(Continued on page 77)

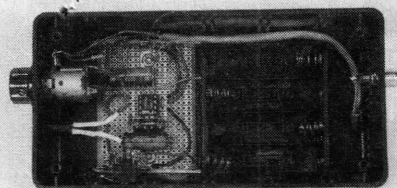


Fig. 3. The Television Hearing Adapter fits into a standard-sized project case. Note the use of terminal strips on the perfboard; that makes it easier to wire the external components to the board after installation in the case.

TV set and your VCR. If you have either a headphone jack, or an RCA-type jack labeled "Audio Out" or "Audio to Hi-Fi", then you are in luck. On a stereo TV, connect to the left channel only. Depending on the type of TV that you have, the audio jack might have a constant-volume output designed for recording regardless of where the TV's volume control is set. Plugging the Television Hearing Adapter into that output jack will give you the advantage of controlling your volume separately from the TV. Imagine the joy of being able to listen to television with family and friends without "blasting" the audio!

Many VCRs have an audio-output jack for playback to another VCR or a stereo system; that can also be used for the Television Hearing Adapter. Using a jack on the TV is better because you don't have the added annoyance of fiddling with another piece of equipment, detracting from the joy of listening to television.

If your TV gear lacks an appropriate audio output, you are not completely out of luck. One alternative is to connect the unit to your TV's loudspeaker through an isolation transformer such as a RadioShack 273-1380. Connect the 8-ohm secondary winding to the speaker and the 1000-ohm winding to J1. If you don't have an audio-output jack on your TV or VCR, and you don't want to make a connection to your TV's loudspeaker, then the only solution is to buy a new TV!

The Television Hearing Adapter is very sensitive. If you are near a radio station, you might experience radio interference; C1 generally eliminates that problem. If you cannot correct it in any other way, use

a metal box instead of a plastic one to house the circuit.

The Television Hearing Adapter has been tested with audio cables up to 24 feet in length. That length should be more than enough for most installations.

Almost any pair of headphones will work well. If you don't own a pair, most electronics stores have a wide selection to choose from. While the author prefers the lightweight types, try more than one model or type—some sound better than others.

Four AA-size batteries will last from six months to a year in normal use if you don't forget to turn the unit off. A smaller unit can be constructed using a 9-volt battery, but at a usable life of only three to four months before needing a replacement battery. If you decide to go that route, the value of R3 should be changed to 4700 ohms. The author's preference leans toward the four-AA battery version as that yields a longer battery life.

The author would be interested in hearing from anyone who builds the Television Hearing Adapter by sending mail in care of the **Electronics Now** editorial office or by e-mail at raygreen@juno.com. Build the Television Hearing Adapter if you have a hearing problem, and it will give you a whole new outlook on life! Ω

GRAVITY WAVES

(continued from page 44)

changes without ruining our signal-to-noise ratio. However, when we consider the size of the wavefronts we'll be encountering, you'll soon see that this method offers no directional selectivity. Gravity waves generated by every object in the Heavens will be affecting the mirrors from every angle, possibly "swamping out" any usable results. Perhaps by arranging a string of mirrors in a certain manner, we could obtain some directional selectivity—assuming that wave propagation isn't beyond light-speed, where just possibly nothing at all may happen to the dimensions of solid objects!

That's why the heterodyne method appears to be the best

bet: It offers excellent directional selectivity, tunable frequency, and tolerance to wavefront size. However, considering the wavelengths we have to deal with, the interferometer will need an extremely long baseline to function. Remember our one-millisecond pulsar? It would produce the shortest wavelength we could expect to see: 186 miles. That means that if we use a one-quarter wave-length laser spacing, our baseline would be at least 46 miles! Don't forget, most gravity waves that we can expect to detect would have much longer wavelengths than that. The most promising method appears difficult to implement by any practical means.

But that does not mean impossible. Please remember that while some of the concepts presented are a shade radical, they are based on current mainstream theory. It is the author's hope that the above discourse has got you thinking and wondering. And who knows, perhaps one of you may actually be able to build a working gravity-wave detector! Ω

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A UNIVERSAL STEREO NOISE FILTER AND COMPRESSOR

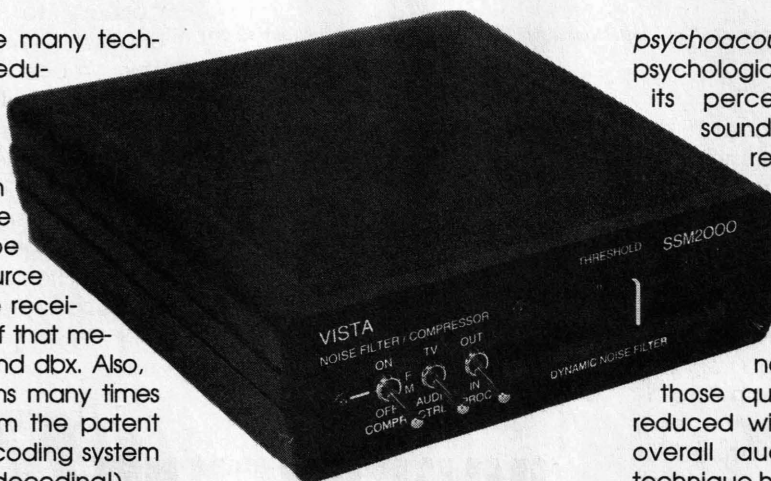
Reduce noise and increase clarity of any audio signal.

While there are many techniques for reducing noise on audio signals, few are universal or "single-ended." Most noise-reduction methods need the original signal to be encoded at the source and decoded at the receiving end; examples of that method include Dolby and dbx. Also, noise-reduction systems many times require a license from the patent owner to use the encoding system (and sometimes the decoding!).

The best solution would be a "universal" noise-reduction system. Such a system should be able to work with any signal without any encoding. What's more, the circuit should be usable by anyone who has the ability to pick up a soldering iron without worry of running afoul of the law.

The good news is that a truly universal noise-reduction system exists. The Universal Noise Reducer presented here has those features that make it a delightful addition to any audio system. Built around a special noise-reduction integrated circuit, it does not require encoded material; the signal-to-noise ratio of any signal can be improved by 25 dB. Any line-level audio source such as tapes, records, FM radio, stereo TV, and satellite signals will benefit. It is also quite effective with CD copies of older analog sources. The system also improves the sound quality of video games and electronic keyboards.

Noise-Reduction Basics. Figure 1 illustrates the frequency and amplitude envelope of a typical audio signal within the bandwidth of an audio amplifier. While the signal might occupy any frequency or



RICHARD PANOSH

amplitude within the amplifier envelope, it does not occupy that area completely at all times. That is especially true during quiet passages or a pause in either the music or speech. During those periods, the large audio bandwidth is not required and noise becomes quite evident. By its nature, noise has a wide bandwidth and is perceived as a hissing sound known as *white noise*. When the audio signal is present, the noise is covered up, or "masked", by the signal. That concept is well known to the field of

psychoacoustics, which studies the psychological aspect of sound and its perception. Whenever a sound is being perceived, it reduces the ability of the listener to hear another sound. During those quiet passages, the noise is "unmasked" and heard as a hiss.

The unmasked white noise that occurs during those quiet passages can be reduced with a reduction of the overall audio bandwidth. That technique has been available for a number of years and is referred to as *Dynamic Noise Reduction* (DNR is the registered trademark of National Semiconductor). An example of that technique was covered in an article that appeared in the July 1994 issue of our sister publication, **Popular Electronics**. That project used a voltage-controlled variable low-pass filter to track the audio-frequency envelope and maintain the required bandwidth to pass the signal information. Since the signal-to-noise ratio is improved with the square root of the bandwidth reduction, a reduction in bandwidth from 35 kHz to 1 kHz can improve the signal-to-noise ratio by

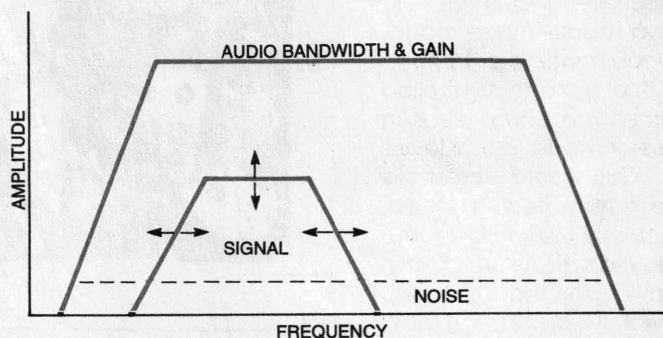


Fig. 1. Noise is inherent in any audio signal. As long as the level of the audio is loud enough, our ears ignore the noise. It's when the music becomes quiet that the noise becomes apparent.

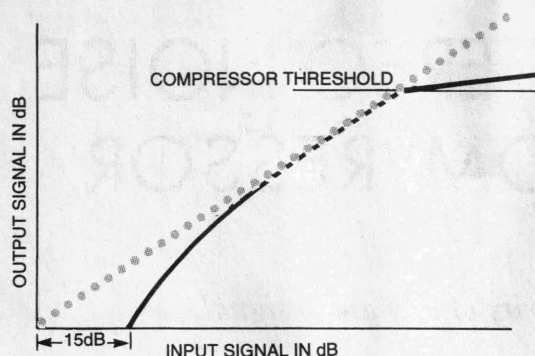


Fig. 2. If louder passages are attenuated and softer passages boosted, a further reduction in noise can occur.

nearly a factor of 6, or about 15 dB.

A second technique that can be used for noise reduction is the use of a variable voltage-controlled amplifier. By controlling the amplifier's gain, the level of the background white noise, or *system noise floor*, can be further attenuated. With this method, signals that are above a certain threshold are passed with a gain of one so that no distortion is added to the program material. However, signals below the threshold are attenuated as shown in Fig. 2; the result is an additional noise reduction of 15 dB. Combining those two techniques results in an improvement of about 25 dB or a factor of nearly 18:1!

A drawback in trying to reduce noise in unencoded signals is that different recording equipment, media, and environments have different requirements. In addition to the methods discussed above, the Universal Noise Reducer uses an automatic adaptive threshold that senses the noise floor of different audio sources. This lets the unit automatically maintain the best possible signal-to-noise level at all times.

An added bonus noise-reduction technique available through the use of the voltage-controlled amplifier circuit is a *compressor*. In the compressor mode, signal levels above an adjustable threshold level will be attenuated. That feature is particularly useful to attenuate commercials that are sometimes much louder than the program material. It can also be used to improve softer sounds below the threshold by attenuating loud sounds, thereby reducing the

dynamic range. The compressor and noise reduction functions are independent and can be used individually or together.

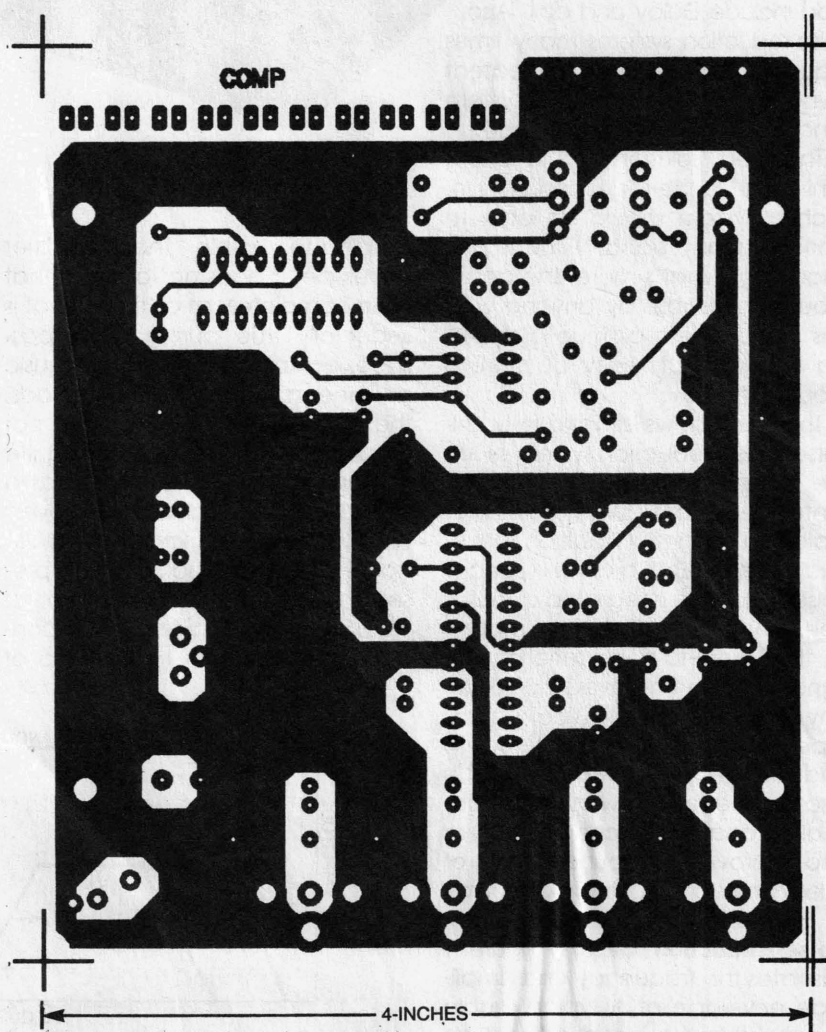
Circuit Description. Follow the schematic diagram shown in Fig. 3

during the following description of the Universal Noise Reducer's circuit.

Power from a 12-volt DC wall transformer is connected to J5. The voltage is regulated to a 9-volt level by IC4. That supply voltage is also divided in half by IC5 to provide a "pseudo ground;" working with op-amps is much easier with a "split" power supply.

Stereo audio signals are input through J1 and J2. Those signals are applied to the heart of the Universal Noise Reducer, IC1. That chip, made by Analog Devices, handles all of the noise-reduction and signal-compression techniques discussed before.

Stereo FM and stereo TV signals use a pilot signal to separate the left and right channels. A tuned passive filter formed by L1, C16, and C13 attenuate any 15-kHz or 19-kHz



Here's the foil pattern for the component side of the Universal Noise Reducer. A large ground plane aids in reducing stray noise pickup.

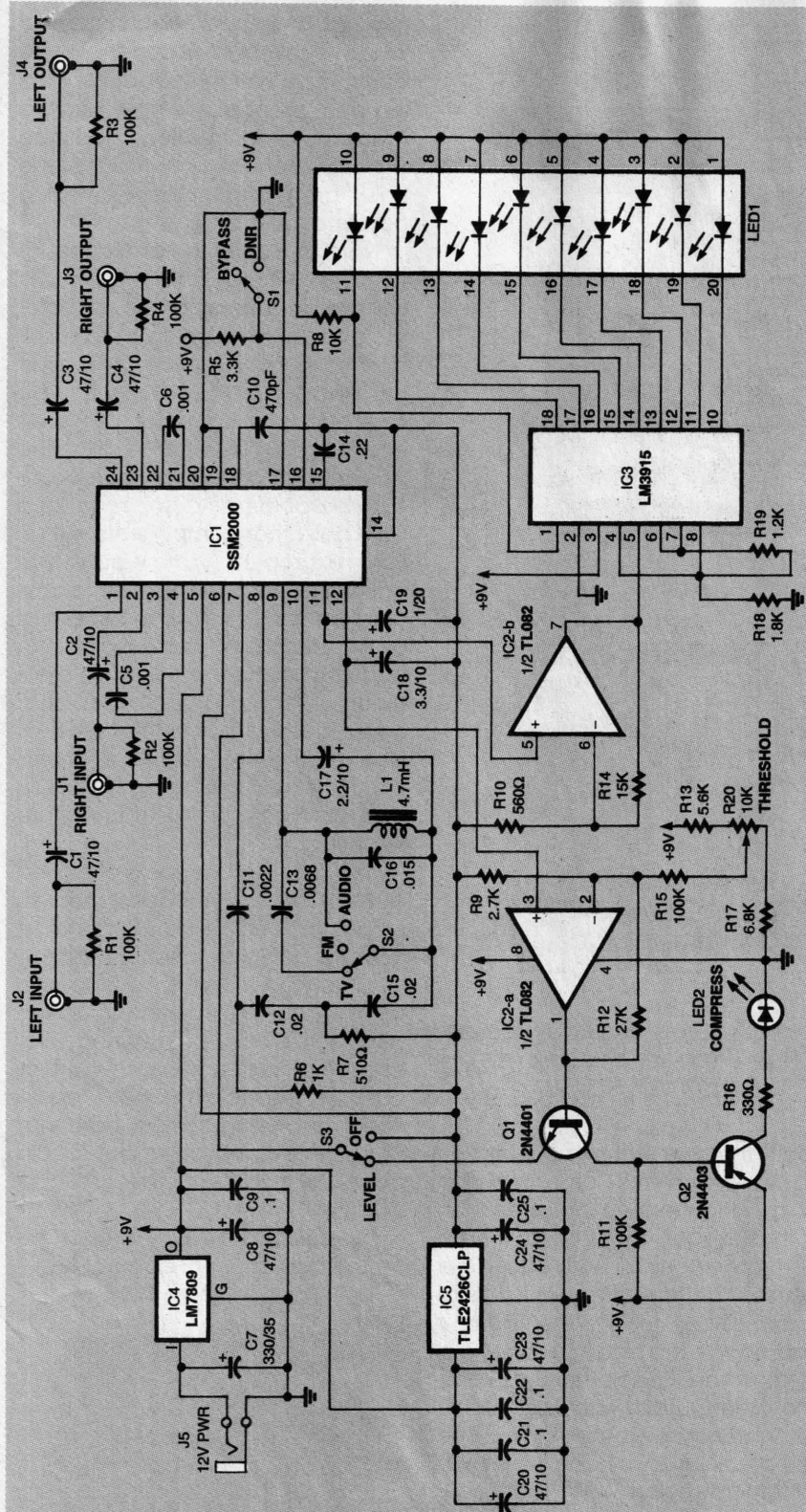


Fig. 3. The Universal Noise Reducer is built around an Analog Devices component that handles all of the noise reduction and compression tasks.

pilot signals if they are present in IC1's control loop. A 15-kHz pilot is used to encode the stereo or SAP (secondary audio program) signals in television; 19 kHz is used for stereo

FM. Switch S2 is used to select filtering for one of those frequencies or no filtering at all. Older audio equipment was more prone to leak the pilot signal into the audio sig-

nal; newer equipment removes the signal better.

Each section of IC2 handles a different function. Op-amp IC2-b is used as a high-gain buffer to isolate pin 11 of IC1—the control voltage for the voltage-tuned low-pass filter's bandwidth. Note that IC1 processes that voltage internally to remove any variations in amplitude. The voltage level present at pin 11 is still a function of signal amplitude but for normal line-level signals it is suitable for an indication of filter bandwidth. The buffered signal is displayed on a bar-graph LED display with IC3; variations in the low-pass filter bandwidth can be tracked. During quiet passages in the audio signal, the low-pass filter should be at the minimum bandwidth of 1 kHz; the first LED should be extinguished. If a pilot signal is leaking into the control loop, the lower LEDs will remain on, indicating that noise reduction is not working as well as it should be.

The other op-amp, IC2-a, closes IC1's compressor loop. It monitors the level of the source-signal detector on pin 12 of IC1. That voltage level is compared to the threshold level set by R20. Levels above the threshold level switch Q1 on. The feedback signal passes through S3 to pin 7 of IC1, which then attenuates the signal until the level falls enough to turn IC1-a and Q1 off. If the compressor feature is not wanted, simply throwing S3 breaks the feedback loop. When Q1 switches on, the voltage drop developed by R11 switches on Q2 and, in turn, LED2, indicating that the compressor is active.

The processed output signal is available on J3 and J4. Switch S1 is used to bypass the Universal Noise Reducer without having to disconnect the unit from the signal path. By switching the unit in and out of the audio path, an "A-B" type of comparison is easily done as to the effectiveness of the various settings of the Universal Noise Reducer.

Construction. Because of the nature of audio circuits, the Universal Noise Reducer is best built on a PC board; less stray noise and hum will be picked up. After all, the pur-

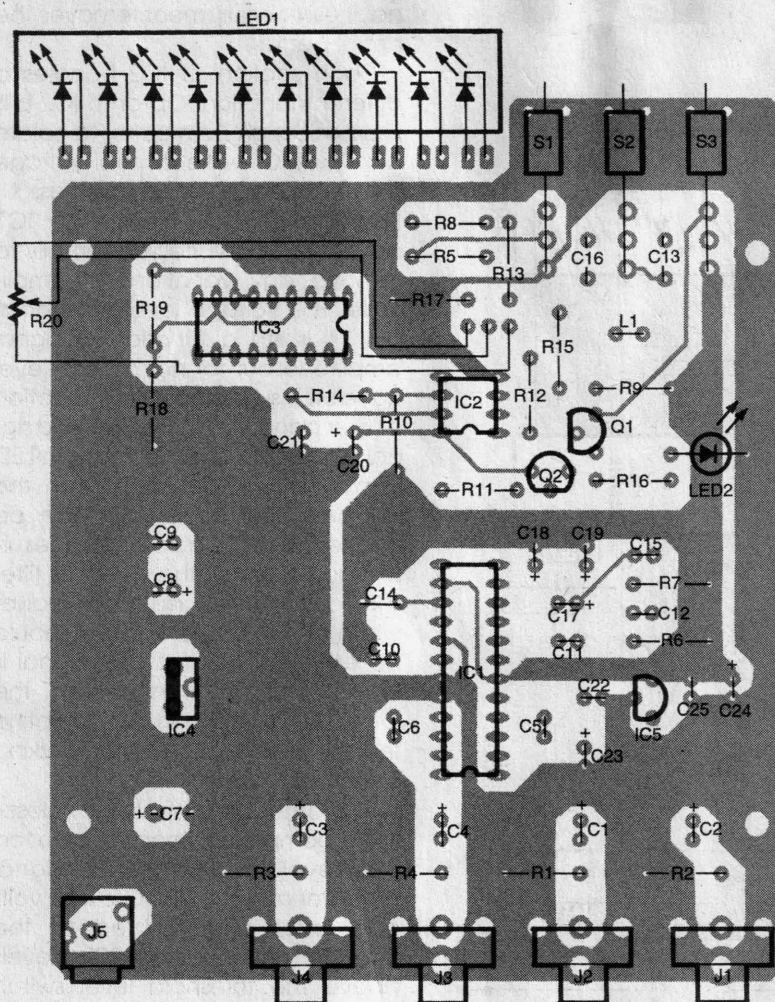


Fig. 4. The Universal Noise Reducer uses a double-sided PC board. If your board doesn't have plated-through holes, don't forget to solder the traces where pads are on both sides of the board. No jumpers are needed; all connections between sides are made at component leads.

pose of the project is to *reduce* noise—not add to it!

Foil patterns for a double-sided layout have been included here for those who wish to etch and drill their own board. If you do not want to etch a board, one is available from the source given in the Parts List.

If you are using a board from the source given in the Parts List or one made from the foil patterns, use the parts-placement diagram shown in Fig. 4. The best order for mounting the components is to start with the smallest parts first; electrostatic-sensitive parts, such as semiconductors, should be mounted last irrespective of their size. If you want to use sockets for the integrated circuits, mount them first, followed by the resistors and capacitors.

Note that if you are using a PC

board that you etched yourself, you will have to make some provision for making connections between both sides of the board where there are traces on both sides. One method (certainly the easiest) is to simply solder to the component lead on the top side as well as the bottom side. Special eyelets have also been used with success in both industry and the hobby. The ultimate solution, of course, is plated-through holes. While that technique is simple to do in an industrial or commercial setting, it is messy, time-consuming, and occasionally dangerous due to the type of chemicals and equipment needed for creating a plated-through hole. The foil patterns do not have any "via" holes (holes that pass a signal from trace to trace between the

two sides without a component lead in them) in their design, so you don't have to worry about soldering short pieces of jumper wires to make those connections. If you are using a purchased board, all of the holes are plated through, making construction a snap.

Note that all of the semiconductors, as well as the electrolytic capacitors, are polarized; double-check their orientation before soldering them in place. Any part installed backwards will be destroyed as soon as power is applied for the first time. It is also possible that an improperly-installed polarized component can take out other components with it. For example, an electrolytic capacitor that is installed backwards might work for a few seconds or minutes; but it can explode or burst into flames, sending an electrical surge into a nearby IC that could destroy that component as well. Also beware of components that look the same, such as IC5, Q1, and Q2. Always check twice before soldering!

The switches and jacks specified in the Parts List are designed to mount directly to the PC board. If you are going to substitute panel-mounted parts, you'll have to wire them up in the traditional fashion with short pieces of insulated wire. Keep the wire lengths as short as possible, especially where audio signals are being carried. Note that S2 is a three-position unit; it should be in the middle of the group of switches.

The leads for LED1 should be bent down at a 90-degree angle

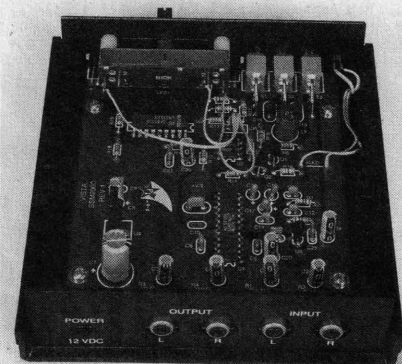


Fig. 5. The Universal Noise Reducer fits easily into a stylish case.

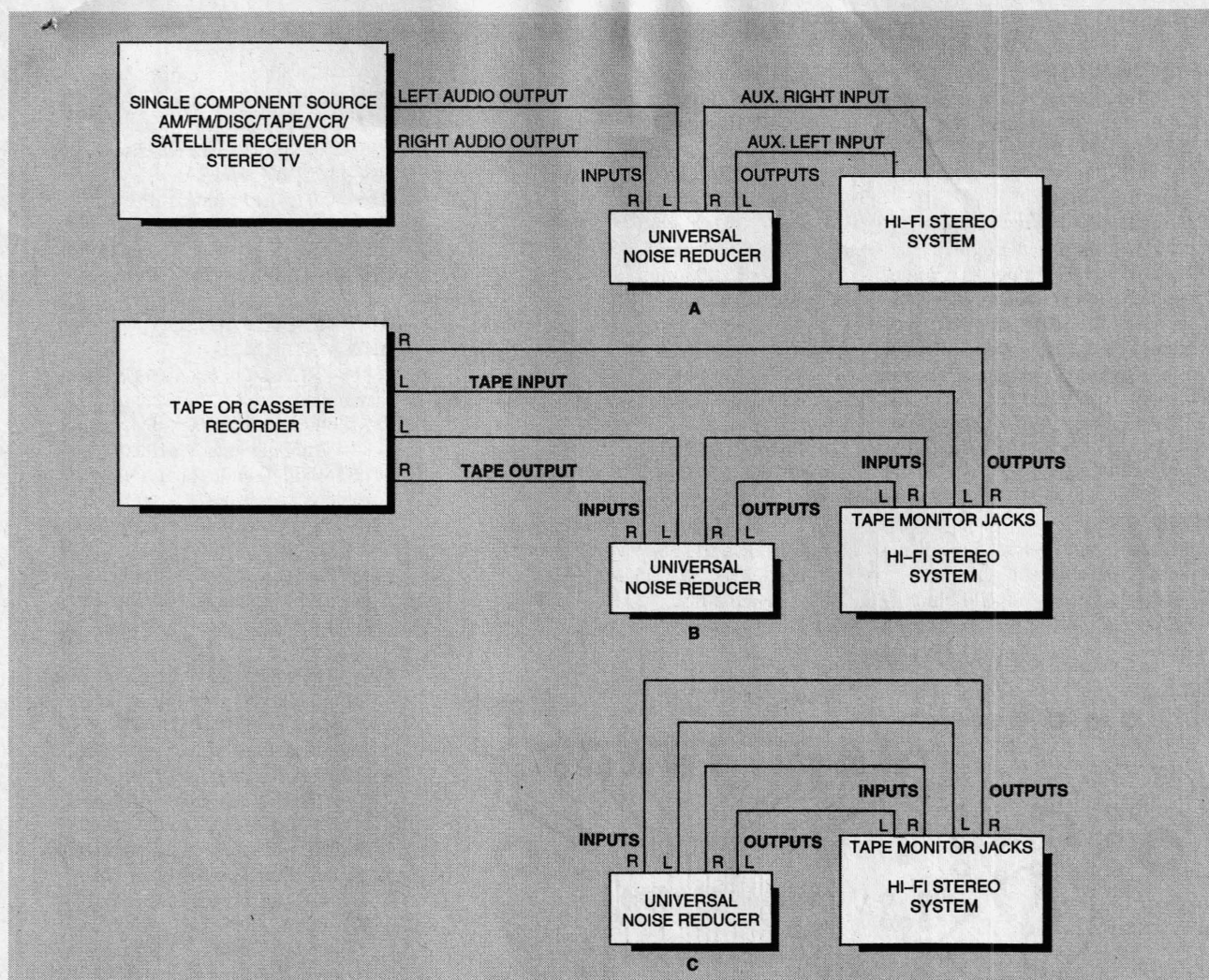


Fig. 6. There are several ways to connect the Universal Noise Reducer to your stereo equipment. It is important to remember that the system works best with an unmodified audio signal. Even if you don't have an amplifier that allows processing equipment to be added to the path, you can always use the tape recording and monitoring jacks.

about 1/4-inch back from the plastic case. A 1/4-inch thick piece of stock can be used as a guide. Check the polarity of the LEDs before bending the leads. A longer (anode) lead will be nearest the corner of the PC board. The display is seated down against the board to obtain the correct height.

When you are finished mounting all of the components, double-check your work for bad solder joints, poor workmanship, and missing, incorrect, or backward components. When you are satisfied with your work, mount the board in a suitable enclosure. The author's prototype, shown in Fig. 5, used a PacTec enclosure. Drill holes as needed for the jacks, switches, displays, and controls. Note that R20 is a slide potentiometer rather than

the traditional rotary style. It is mounted above LED1 and connected to the board by short lengths of insulated wire.

Once the Universal Noise Reducer is mounted in its case, it is ready for testing and use.

Set-Up and Operation. The Universal Noise Reducer can be hooked up in a couple of ways. The most obvious way is shown in Fig. 6A, where the audio source is connected to J1 and J2, with J3 and J4 connected to the amplifier's inputs. Wherever possible, the unit should always be placed in the audio path after any preamplifiers and before the volume control or tone controls. That way, the signals will be of a uniform level and provide a flat-frequency response. Many modern amplifiers

have separate preamplifier output jacks for just such a use.

If you don't have such a rig, the Universal Noise Reducer can be placed in the tape-loop or signal-processing loop; see Fig. 6B. Again, that will provide flat-frequency signals with levels in the range of 100 millivolts to 1 volt rms. If you have an equalizer in the same loop, it is important to locate the Universal Noise Reducer *before* the equalizer. Again, the unit works best with a flat bandwidth before you do any equalizing. If you don't have a tape unit, simply connect the Universal Noise Reducer in place of one as shown in Fig. 6C. By monitoring the tape playback, the unit becomes a part of the audio path on even the most basic stereo equipment as long as tape input and output

PARTS LIST FOR THE UNIVERSAL NOISE REDUCER

SEMICONDUCTORS

IC1—SSM2000 stereo noise attenuator/compressor, integrated circuit (Analog Devices)
 IC2—TL082 operational amplifier, integrated circuit
 IC3—LM3915 bar-graph display generator, integrated circuit
 IC4—LM7809 9-volt positive voltage regulator, integrated circuit
 IC5—TLE2426CLP split power supply ground converter, integrated circuit
 LED1—Light-emitting diode bargraph display, 10-segment
 LED2—Light-emitting diode, green
 Q1—2N4401 silicon transistor, NPN
 Q2—2N4403 silicon transistor, PNP

RESISTORS

(All resistors are 1/4-watt, 5% units unless otherwise noted.)
 R1—R4, R11, R15—100,000-ohm

R5—3300-ohm
 R6—1000-ohm
 R7—510-ohm
 R8—10,000-ohm
 R9—2700-ohm
 R10—560-ohm
 R12—27,000-ohm
 R13—5600-ohm
 R14—15,000-ohm
 R16—330-ohm
 R17—6800-ohm
 R18—1800-ohm
 R19—1200-ohm
 R20—10,000-ohm, linear potentiometer

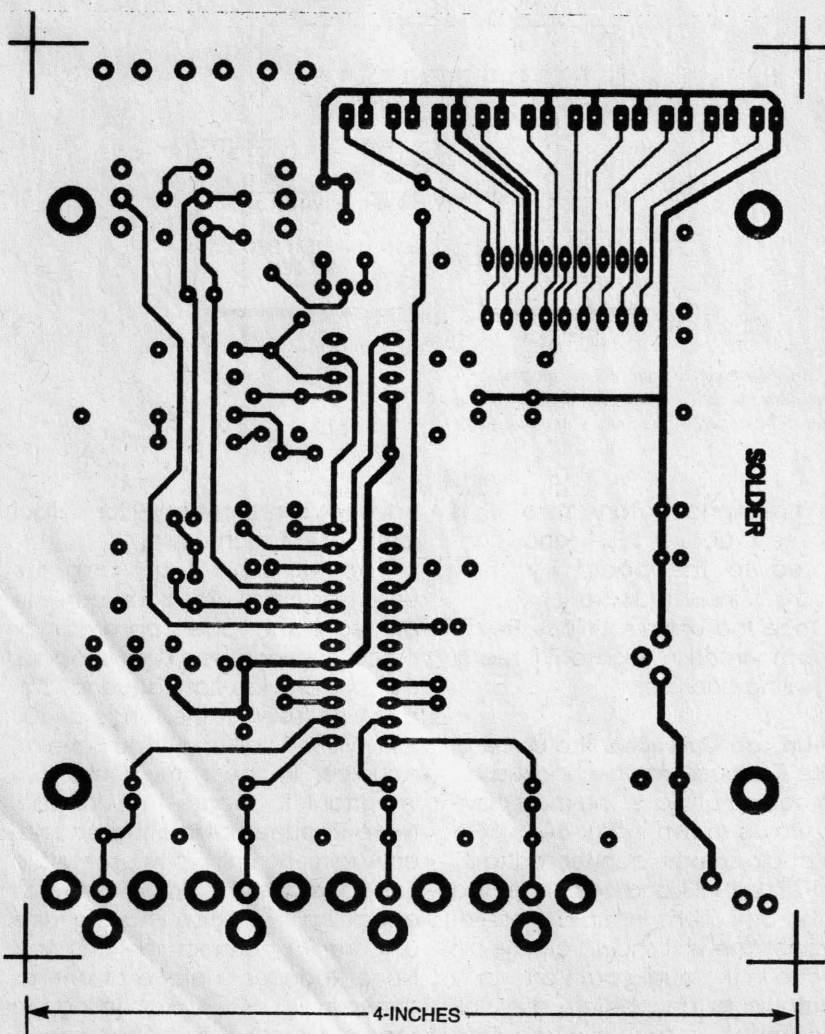
CAPACITORS

C1—C4, C8, C20, C23, C24—47- μ F, 10-WVDC, electrolytic
 C5, C6—0.001- μ F, ceramic-disc
 C7—330- μ F, 35-WVDC, electrolytic
 C9, C21, C22, C25—0.1- μ F, ceramic-disc

C10—470-pF, ceramic-disc
 C11—0.0022- μ F, ceramic-disc
 C12, C15—0.02- μ F, ceramic-disc
 C13—0.0068- μ F, ceramic-disc
 C14—0.22- μ F, ceramic-disc
 C16—0.015- μ F, ceramic-disc
 C17—2.2- μ F, 10-WVDC, electrolytic
 C18—3.3- μ F, 10-WVDC, electrolytic
 C19—1- μ F, 10-WVDC, electrolytic

ADDITIONAL PARTS AND MATERIALS

J1—J4—RCA-style phono jack, PC-mounted
 J5—Co-axial power jack, PC-mounted
 L1—4.7mH inductor, shielded
 S1, S3—Single-pole, double-throw switch
 S2—Single-pole, double-throw, center-off switch
 Case (PacTec CM125 or similar), 12-volt DC wall adapter, IC sockets, knob for R20, wire, hardware, etc.



Here's the foil pattern for the solder side of the Universal noise Reducer. All connections between the two sides are done at component pads, making it easier to build the unit without using plated-through holes.

Note: The following items are available from Vista, PO Box 142 Bolingbrook, IL 60440-1532; Tel: 630-378-5534: Assembled and tested unit (SSM-2000-ASSEM), \$125.00; Complete kit of all parts and case (SSM-2000-KIT), \$95.00; etched, drilled, silk-screened PC board with plated-through holes (SSM-2000-BOARD), \$18.00; IC5, \$3.90; IC1, \$18.00. All mail orders will be shipped by first class US mail with a shipping/handling charge of \$3.00. All telephone orders will be shipped by priority mail with a shipping/handling charge of \$6.00. Illinois residence must include 7.5% sales tax. Visa, Mastercard, and American Express credit cards are accepted.

jacks are available.

With a 12-volt power source connected to J5, throw S1 to BYPASS, S2 to AUDIO, and S3 to OFF. Unmodified audio should be heard. Switch S1 to DNR and a reduction in noise should be heard. Try using audio sources that have different gain levels, such as a quiet passage of a piece of classical music. If you are using a television or radio as an audio source, you might have to set S2 to the appropriate mode as mentioned before in case the carrier signal is leaking into the audio path. Finally, throw S3 to test the compressor; note that compression takes place at different levels by varying R20. The level of compression should be seen on LED1. Ω

Pour an Antenna for X-Band

20-20 vision for your microwave system!

by John M. Franke WA4WDL

Ever heard of anyone pouring an antenna? This idea isn't as silly as it sounds, and it's quite easy.

What I am referring to is casting lenses—microwave lenses. It turns out that optical lenses above 3 GHz actually help correct aberrations in the wavefront coming out of a horn antenna, thereby reducing sidelobes. A lens collects and collimates the energy coming out of the horn which results in an increase in gain.

Some time ago, Bill Hoisington K1CLL described a simple X-band (10 GHz) crystal set using a 1N23 detector and a hemisphere of wax as the antenna. (See "The World of X-Band" by Bill Hoisington K1CLL, 73, March 1975.) He cast ordinary kitchen paraffin to make several short focal length lenses about 0.64cm and 12.7cm in diameter. Paraffin is not optically transparent, but microwaves do not know that. The wax is very clear to microwave energy. Also, the lens surface does not have to be optically smooth. As long as the imperfections are less than an eighth wavelength (0.142 inches), there should not be any noticeable effects. Lenses are much more tolerant of surface errors than are reflector or "dish" antennas. A surface error on a reflector is doubled upon reflection. For example, a tenth-wave bump or dent produces a fifth-wave distortion in the wavefront. A microwave lens with an index of refraction of 1.5 and having a tenth-wave bump or dent produces a wavefront distortion of one-tenth times 1.5 minus 1, or one twentieth wave. In other words, the distortion is less than the physical defect.

Designing and Building the Lens

I can't go into detailed lens design, but I can relate some personal experience in casting lenses. After all, most of us started in microwaves by using dish or horn antennas we found through surplus sales or ham-fests. Then, as special needs arose, we designed and constructed custom antennas. The majority of dish and horn antennas used by amateurs fall into the found or purchased category. Few of us actually build them. I have built a couple of special purpose dishes, but will opt for an already-built one any chance I get—I'm basically lazy. (See "The Amazing Cylindrabola" by John M. Franke WA4WDL, 73, September 1983.)

Several years ago I came across a large glass lens at an antique flea market. I've been keeping it for use as an optical receiver anten-

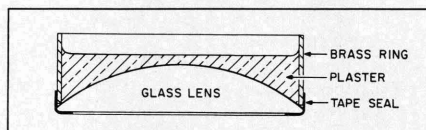


Figure 1. Forming the plaster mold.

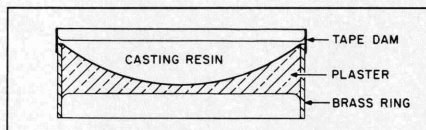


Figure 2. Casting the replica lens.

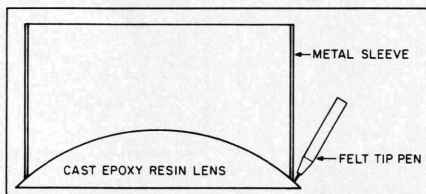


Figure 3. Marking the cast lens for edging. Note that the marking sleeve or cylinder must be perpendicular to this rear surface of the lens.

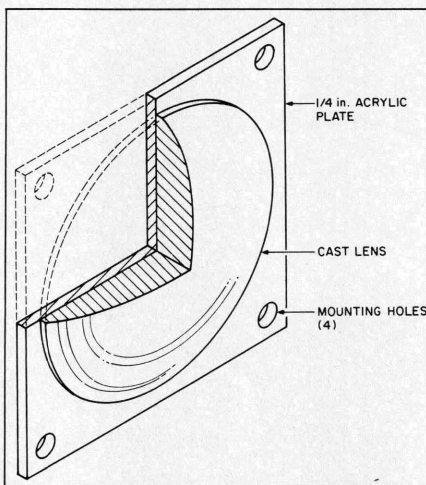


Figure 4. Cast lens epoxied to acrylic support plate.

na. It is 6 inches in diameter, about 1 1/2 inches thick, and quite heavy. It has a focal length of about 6 inches. The lens is flat on one surface and convex on the other, so it is known as a "planoconvex." Such a large fat lens is poor for optical imaging, but it's a great light collector. However, I needed two antennas. So, I decided to try to cast epoxy resin replicas of the glass lens.

In my junk box I found a couple of brass rings 6 inches in diameter and 2 inches tall. If I had not had the rings, I could have made suitable rings or sleeves by forming strips of aluminum flashing into a band and taping it to hold its shape. I placed the glass lens on the first ring, curved surface down, then taped the glass to the metal edge with plastic electrical tape, forming a watertight seal. (See Figure 1.)

Next, I mixed plaster, poured it into the mold, and let it harden for several hours. (See Figure 2.) Once the plaster had hardened, I removed the tape, separated the glass lens from the first ring, and taped it to the second ring. Then I poured the second mold and let it harden. (I let both plaster castings dry thoroughly for two days before doing any more work.)

In Photo A you can see several screws protruding through the sides of the brass rings into the plaster. The screws are there to plug existing threaded holes in the rings. They have no other real function, but they may be helpful in holding the ring and plaster mold together.

Once the plaster had dried, it was time to cast the lenses. One mold would have been sufficient, but I wanted to mix the epoxy resin only once. So, I made two molds to have added assurance against having to throw out the mixed resin if I dropped or chipped a mold after the resin was mixed. I placed the molds on a level surface with their smooth cast surfaces up. (If the molds are not level, the rear surfaces of the lenses will be wedged.) A strip of plastic electrical tape was used to form a straight wall lip at the top of each mold. I mixed the hobby casting resin using the least amount of catalyst recommended. This lengthens the curing time, but reduces the

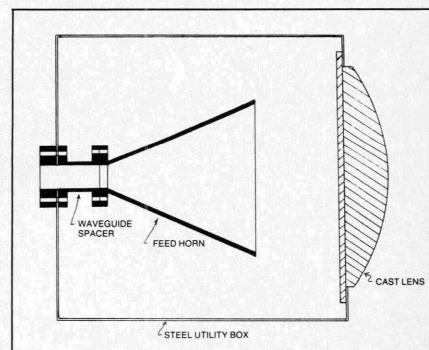


Figure 5. Completed antenna mounted in a 6-inch cube utility box.

chances of uneven curing and cracking due to rapid heat build-up. Next, I poured the resin and allowed it to cure overnight. The curing site should be outside to eliminate any problems with the smelly fumes.

The next day, I separated the cast lenses from their molds. The molds survived and could be used again. The edges of the lenses were rough and chipped, so I decided to trim or "edge" the lenses. I used a cylinder with an outside diameter of 5 inches—the hood from an old oscilloscope—to mark the lenses. (See Figure 3.) The actual edging was done with a disk sander and only took a couple of minutes for each lens. After the edging, the next task was to mount the lenses. Normally, lenses are mounted in a finely-machined barrel with a threaded retaining ring. I can operate a lathe and chase threads, but let's be reasonable! Instead, I epoxied a 5½-inch-square piece of ¼-inch acrylic to the flat surface of each of the epoxy lenses. (See Figure 4.) Then I drilled four mounting holes in each acrylic square.

Light sanding of each lens, followed by a couple of coats of clear acrylic spray, finished the lenses. The sanding is for two reasons. First, although the lenses are of poor optical quality, they can still produce a very hot spot when aimed at the sun, unless their surfaces receive a ground scattering finish. Second, they look better with an even, frosted appearance.

Testing the Lens

One nice thing about small aperture microwave antennas is that you can set up a test range in a short distance. For example, with a small horn source antenna and the cast lenses, a distance of 20 feet is more than adequate. For a source, I used a surplus 2K25 klystron that had its cavity stretched to operate on 10.4 GHz. (See "A Complete X-Band Transmitter" by Stirling M. Olberg W1SNN, 73, August 1978.) People may shun vacuum tubes, but I can only respect a tube that, like me, is over 40 years old and still working. Besides, have you priced a Gunn diode lately? I would very much like to hear from anyone with a supply of Solfan Gunn diode units. Anyway, the 2K25 puts out about 25mW when used with an old surplus klystron power supply that I got at a hamfest many years ago. The receiving portion of my antenna range is a 0–40 dB attenuator and a tunable detector mount. For metering, I now use a circuit published by Chuck Houghton WB6IGP. (See "Microwave Test Equipment for 10 GHz" by Chuck L. Houghton WB6IGP, 73, October 1988) I did

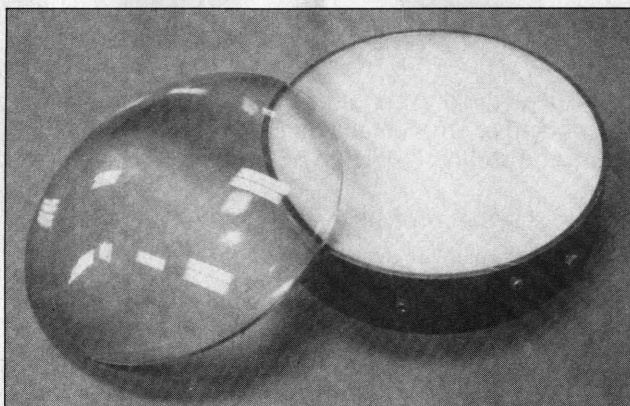


Photo A. Original glass lens and plaster mold.

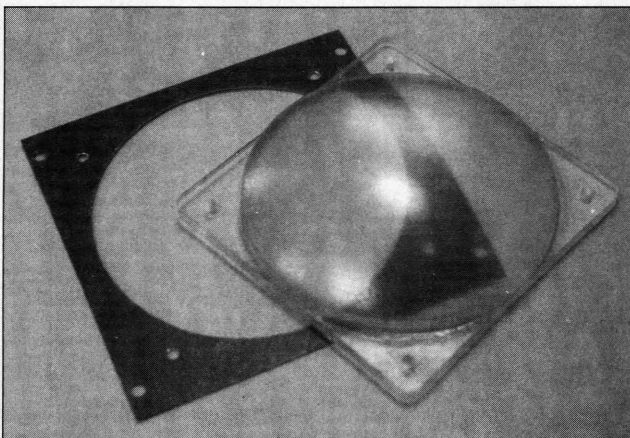


Photo B. Cast lens epoxied to acrylic plate and lens mounting plate.

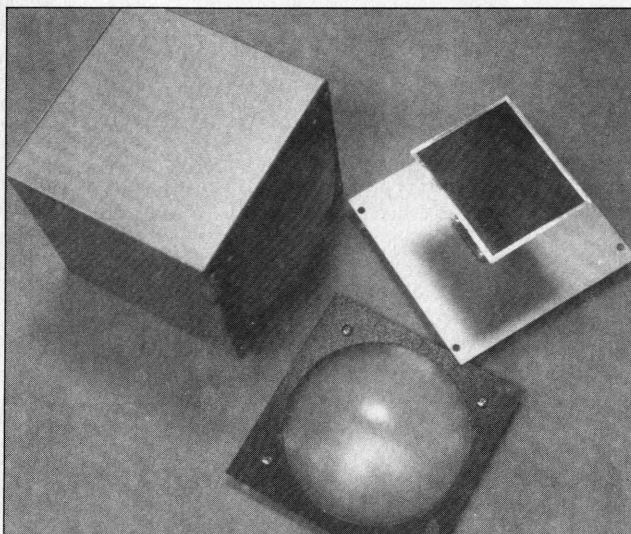


Photo C. Mounted lens, mounted horn and complete antenna.

have to reverse the polarity of the detector diode to get it to work, probably due to a misprint. His circuit is very compact and very sensitive. You cannot go wrong reading his many articles, and he now has a superb column, "Above and Beyond," in 73 every month.

I placed a standard gain horn on both the transmitter source and the receiving setup, and tuned the detector mount for maximum response. Then I set the S-meter to midscale by adjusting the variable attenuator. The attenuator setting at this point became my reference reading. Substituting other antennas on

the receiver setup, the attenuator is changed to maintain a midscale meter deflection. The reference reading is subtracted from the new setting. The result is added to the gain of the reference antenna to yield the gain of the new antenna.

Sound more complicated than it actually is? Say you have a reference antenna of 16 dB gain and with it you get a reference attenuator setting (reference reading for midscale meter deflection) of 8.5 dB. Putting your test antenna on the receiving setup, you get a midscale meter deflection with the attenuator changed to a setting of 12 dB. Then, subtracting the reference reading ($12 - 8.5 = 3.5$ dB), you find that the test antenna has a gain 3.5 dB greater than the reference. You have to attenuate the signal from the test antenna to make it equal to the reference signal. The gain of the test antenna is the difference added to the gain of the reference antenna ($16 + 3.5 = 19.5$ dB). If the midscale deflection had required an attenuator setting of 4 dB for the test antenna, then subtracting the reference reading ($4 - 8.5 = -4.5$ dB) tells you that the test antenna has a gain of 4.5 dB less than the reference, or a net gain of $16 - 4.5 = 11.5$ dB.

You do not need to know the absolute gain of an antenna to use this technique. As long as you use the same antenna as a reference, you can measure the relative gain of your other antennas to it. After all, we are more interested in reducing losses or improving gains than we are in knowing what the absolute levels are.

Back to the lenses. The lenses, having a focal ratio (or F#) of about 1, should be fed with some sort of small horn antenna. I used the only one I had at the time—a Microwave Associates metalized plastic 17 dB gain horn. The addition of the epoxy lens increased the gain by 4.5 dB, or a factor of 2.8. With a unit at each end of a path, the net system gain is 9 dB, allowing the usable separation to be increased by 2.8 for equal signal strength. I

was able to mount the lenses on one face of a 6-inch cube steel utility box and mount the horn with a spacer on the opposite face of the box. The result is a compact, easy-to-handle unit that can be readily mounted on a tripod for field work, or on a tower for point-to-point applications. (See Figure 5.)

I am now looking for a larger lens to cast, or perhaps I can find a large bowl to serve as a mold. **73**

Contact John M. Franke WA4WDL at 23 Parkwood Drive, Apt. 201, Yorktown VA 23693.

An Inexpensive 10 GHz Dish System

The plumber's delight!

by Jerry Jensen WT0W

This design for a 10 GHz dish and feed system was born of necessity. There are few sources of microwave components here in the Midwest, so construction projects have to use available non-microwave parts. The only microwave part used for this system is one WR90 waveguide flange. Even that could be made of scrap material if you can't locate one easily.

The dish is an old 18" surplus HBO dish left over from the days when premium channels were broadcast. It has a focal ratio of 0.4, providing a focal length of about 8.1". The dish I found has a 1" hole, conveniently located at its center, that permits the 10 GHz source to be mounted behind the dish and fed through it to a modified "penny feed." A chassis punch or drill could be used to make the hole on a solid dish. One of the old "flying saucer" dish snow sleds would also work if you modify the feed's length to meet the new focal ratio.

The feed system starts out as a 12" piece of $\frac{3}{4}$ " copper pipe. Anneal one end of the pipe by heating about 2" of the end until the surface looks very clean (don't get it red hot), then plunging the hot end into cold water. This will make the annealed end of the pipe very soft and easy to work. Place about $\frac{1}{2}$ " of the end in a vise and squeeze the pipe with the vise until the pipe is an ellipse about $\frac{1}{2}$ " wide. Next, rotate the pipe 90 degrees and do the same thing, but stop at 1", at right angles to the first squeeze. By repeating the operation a few times, you will end up with an end on the pipe that fits the WR90 flange. Pliers and careful hammer tapping (with a sharp right angle piece of steel held inside the pipe) will make the corners of the squeezed pipe perfect. Be careful not to strike the copper hard enough to stretch it or you will end up with too large an opening.

Once the end of the pipe matches the flange, force a $\frac{3}{4}$ " outside pipe coupling (just a copper tube that fits over the joint of two $\frac{3}{4}$ " pipes) down the pipe to as near the flange end as possible. This will act as a spacer for the next plumbing piece, a 1" male pipe threaded to a 1" copper pipe adapter. Put the adapter over the pipe cou-

pling with the male threaded end away from the waveguide flange. Some adapters fit over the coupling tightly, others are a bit sloppy. A layer of clean copper wire can be wrapped around the coupling, or a single turn used at the end(s) for spacers as needed.

Before the assembly is soldered, mount a 1" galvanized pipe flange on the back of the dish over the 1" feed hole in the center. Screw the copper adapter about halfway into the flange and measure the length of the pipe sticking into the dish. Check to be sure that the end of the pipe goes into the dish at least far enough to be within an inch of the focal point. If everything checks out, solder the waveguide flange to the pipe and the pipe coupling and adapter assembly. If you get any solder inside the end of the waveguide at the flange, use a small file to remove it.

Screw the feed assembly about halfway into the pipe flange, mark and then cut the copper pipe to make it about 1 $\frac{1}{2}$ " short of the focal point.

At this point there is an option. If you want to be able to tear down the system into its most portable state for backpacking or such, you can make the splash feed removable. This will allow you to pull the feed system out of the dish. The feed system and splash plate can then be packed inside the dish.

The splash plate assembly is made from a male $\frac{3}{4}$ " pipe threaded to a $\frac{3}{4}$ " copper pipe adapter, a PVC (plastic) $\frac{3}{4}$ " female pipe threaded to a pipe adapter, and a $3\frac{1}{4}$ " copper disk. The PVC adapter is epoxied to the center of the copper disk. This is then screwed on to the pipe thread of the $\frac{3}{4}$ " copper adapter and the disk. For a completely portable system, drill and tap the copper adapter for a set screw to hold it to the feed system (an extra nut soldered to the outside might be wise). For fixed operation, solder the copper adapter to the radiating end of the feed system.

Tune Up

Tune up is simple: The radiating end of the feed has to be set to the dish's focal point. Then

the separation between the end of the waveguide and the splash plate is adjusted for maximum radiation.

You will need some sort of power or signal strength measuring device to make these adjustments. It can be as simple as an unpowered Solfan unit. Just measure the mixer diode current of the unpowered unit when it is placed a few feet away from the dish with a sensitive current meter. Another Gunn transceiver could also be used to peak the signals.

The feed system can be adjusted by screwing the assembly into or out of the dish. The splash plate can be adjusted by screwing the PVC adapter further on or off the $\frac{3}{4}$ " pipe thread. These adjustments interact and should be made to optimize the gain of the system.

If you have built the portable system with set screws, be very careful not to distort the $\frac{3}{4}$ " copper pipe waveguide when you tighten the screws. If everything has been done correctly, the polarization of the output will be within a few degrees of the original source. If small dents or distortions are introduced into the pipe, the polarization of the output will rotate.

An 8-32 screw can be inserted into the center of the copper disk and adjusted to scatter the RF energy better as it hits the disk. The copper disk can be made from a piece of single or double-sided PC board. The PVC adapter could be machined from a better microwave plastic for lower losses. During the machining, a dielectric lens structure could be formed to optimize the scatter from the disk.

This disk and feed system may not be perfect, but it is a usable system that can be made from hardware store plumbing parts. Total cost, without the dish, should be less than \$5. The feed illumination is good, and there is very little measurable radiation escaping the dish. Measurements taken across the dish (built with the dimensions shown) show a reasonable illumination pattern. **73**

Contact Arthur J. (Jerry) Jensen WT0W at 10900 Ewing Ave. S., Bloomington MN 55431.

Parts List

- 1 12" length of $\frac{3}{4}$ " copper water pipe
- 1 $\frac{3}{4}$ " male pipe thread to $\frac{3}{4}$ " pipe adapter
- 1 $\frac{3}{4}$ " female-female pipe coupling
- 1 $\frac{3}{4}$ " PVC female pipe thread to PVC pipe adapter
- 1 1" copper male pipe thread to pipe adapter
- 1 1" galvanized iron female pipe flange
- 1 WR90 waveguide flange
- 1 $3\frac{1}{4}$ " copper brass disk

Note: 18" and 24" dishes are available from Satellite City, 2663 County Road I, Moundview MN; (612) 786-4475. Costs are about \$20 and \$30 respectively.

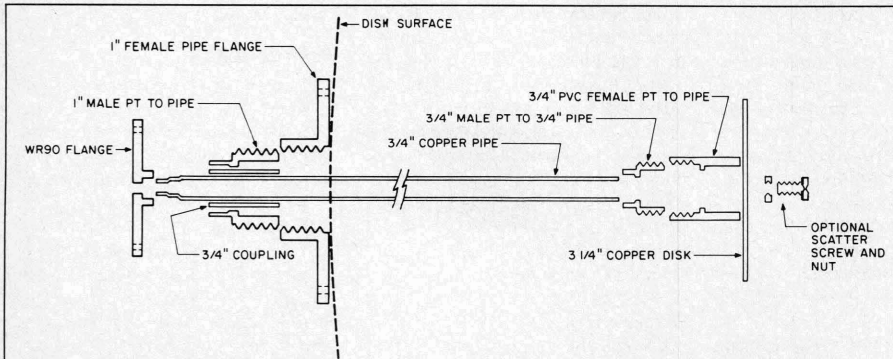


Figure 1. Assembly of the 10 GHz dish and feed.

1220 MHz— Use It Or Lose It!

—simple gear you can build and enjoy

This article, for the genuine amateur radio and electronics experimenter, details the construction of a crystal-controlled exciter, an oscillator, a tuned diode detector, and an eleven-element beam covering the amateur band of 1215 to 1300 MHz. In these days of "store-bought" rigs on lower frequencies, there is a particular fascination in working

with circuits which cannot be found down on "Radio Row."

This particular amateur band is about the last one, going up in frequency, on which a Yagi beam is still useful. At 2400 MHz, almost any kind of a dish is better, but, here on 1220, long Yagis will still give plenty of gain—like close to 20 dB. And, look at the low cost! A piece of

for 1220 MHz, can be made out of copperclad and brass sheet, although spring copper or beryllium copper make better sliding contact springs, of course.

First there's the $\frac{1}{4}$ -wave open-ended circuit. I have made some of these operate on 1220, but you're right on the end of nothing as far as a decent tuned circuit goes. Fig. 1 shows a tuned diode detector circuit which is good up to about 800 MHz. Actually, this one is about all through at 1000 MHz, so don't try to use it for 1220 unless you want to just play around.

Then there's the $\frac{1}{2}$ -wave "double-short" circuit. This one, shown in Fig. 3(a), works fine on 1220. I've even used it right up to and including X-band (10,000 MHz), although there it really needs a $\frac{1}{4}$ -inch i.d. cavity with a one-eighth-inch center conductor, but that's another story.

As in Fig. 3(a), the ever-useful copperclad forms the baseboard, with another piece one inch wide for L1 fastened down at both ends with insulation. I made the sliding shorts out of brass sheet, although spring copper is better. It is also supposed to work better if silver-plated (and, of course, it will, at least a little). I have not

PVC tubing, a chunk of no. 12 bare copper wire cut into little pieces, some Elmer's glue, and you've got it.

Of course, finding the lengths and spacings of each one of those elements is very important, but I've already done that part for you.

The tuned circuit itself,

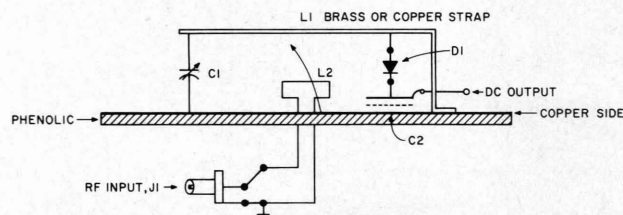


Fig. 1. $\frac{1}{4}$ -wave circuit. Do not use above 800 MHz.

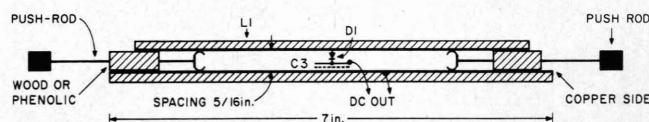


Fig. 3(a). Completed diode detector. Tunes 1000 to 1400 MHz. C3—see Fig. 3(b).

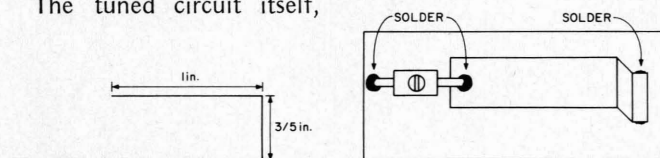


Fig. 2(a).

Fig. 2(b).

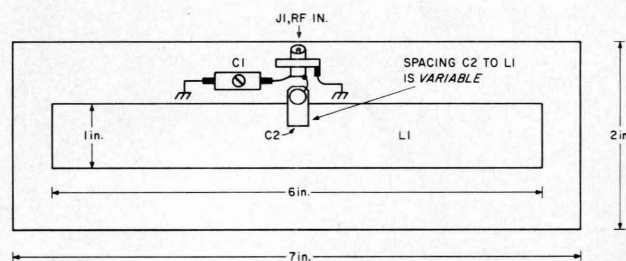


Fig. 3(b). Top view. C2 is variable (about $\frac{1}{2}$ pF). J1 is a phor jack. C3 is 1 inch square, insulated from baseboard. Use nylon bolt to hold C3 to base.



Fig. 4. Sliding shorts.

silver-plated mine yet. Just be sure to make the brass shine.

You will have to put a little effort into making the sliders, as shown in Fig. 4. I was doubtful at first about making them myself, but, after the first few you make, it's not so hard. They do work, if you handle them with care. In Fig. 3(a) I show handles on the push rods, but if you leave these off or make them readily detachable, you can take the sliders out quite easily and adjust them for a good fit, taking care to make the pressure on each finger as nearly equal as you can.

Fig. 4 shows the flat cut, with the fingers, before bending. Fig. 5(a) shows the finished shape, after bending, with push rod soldered in place. Fig. 5(b) shows details of the contacting surfaces bent for best contact operation.

Now that you have a tuned diode detector, you need something to detect, and the oscillator of Fig. 6 will give you just that, although you may wish to proceed directly to the crystal-controlled exciter. Figs. 6(b) and 6(c) give further details on the oscillator.

Good transistors begin to

be a problem for 1220 MHz. I used the old standby 918, which must be at least ten years old by now. I found that tuning on 1220 is remarkably sharp as soon as you get things running at all well. An oscillator is a lot of fun and quite useful for certain purposes, such as tuning up beams (especially when followed by an amplifier for more power), isolation, and modulation. The crystal-controlled exciter is, of course, the unit to use for serious work on the air.

In order to make the multiplier tuning simpler and a sure thing, I decided to stick to doubling in the UHF and microwave stages. If you have had any experience with frequency multiplying at UHF and up into microwaves, you will be acquainted with the fact that the times-two harmonic (doubling) is always the most powerful.

You can start on 50.9 MHz, as in Fig. 7, triple to 152.7, double to 305.4, double to 610.8, and, finally, double to 1221.6 MHz in the amateur band. Note that tripling from 50 to 150 is not too hard, but from then on

up, you should double only for the simplest and surest results. Of course, if you want to, using very high Q circuits and very careful attention to wavemeters, calibrated tuned detectors, and well-calibrated microwave receivers, you can do almost any amount of multiplying. For instance, I worked with a large company in New Hampshire in the late 1960s which was building X-band equipment for the ill-fated F-111 jet fighter. Someone had designed a little unit which was supposed to multiply by *ten*! At X-band! Needless to say, I advised elimination of that unit and came up with a Gunn diode oscillator on X-band. They had been using several X-band spectrum analyzers to find that elusive little tenth harmonic. So stick to doubling; you'll have fewer grey hairs!

You can start in this exciter with almost any good VHF transistor as the 50.9 MHz oscillator, using the well-tried and proven circuit of Fig. 7. Though tripling to 152.7 isn't too difficult, you will need to make up or borrow some VHF and UHF absorption wavemeters. Do not use a receiver to determine which harmonic you are on. Better are the tuned diode types shown in Fig. 1.

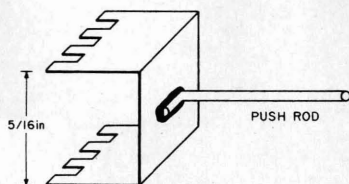


Fig. 5(a).

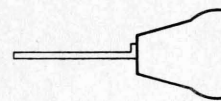


Fig. 5(b). Bending detail of shorts.

They are also excellent for checking AM modulation with a small af amplifier and padded earphones, so you can hear your own modulation as it sounds on the air.

You will have enough to do getting a microwave rig and beams running for interesting DX tests without going to other than AM modulation, at least to start with, so be happy with it. It also works very well and gives you the same or better DX.

The doubler from 152.7 to 305.4 MHz is still easy enough, if you have some experience and frequency calibration available to help you. For the next doubler to 610.8 MHz, I had to use the old reliable 918, and, just to be sure of getting the doubling frequency harmonic on 610.8, I made up a lecher line quickie, as in Fig. 9. This one is really indispensable for checking frequencies over 1000 MHz. Couple J1 to the oscillator or stage under test with a short piece of RG-58/U cable, and observe peaks and/or dips as shown

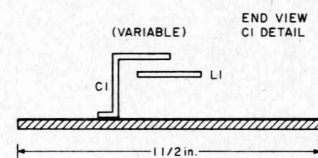


Fig. 6(c). Collector tap at $\frac{1}{2}$ " from end of L1.

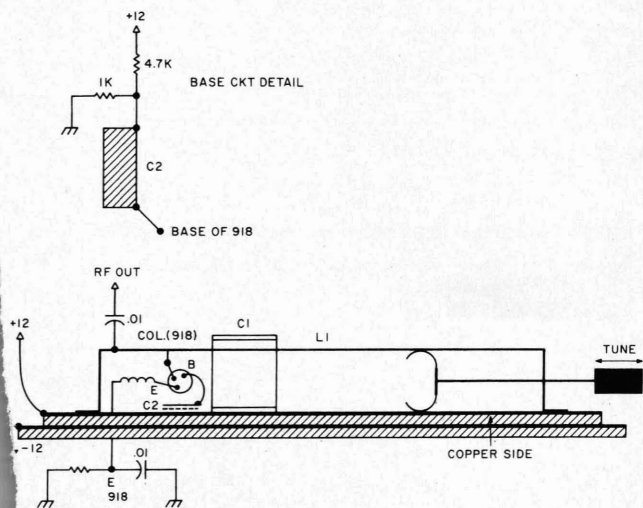


Fig. 6(a). Oscillator, 1220 MHz.

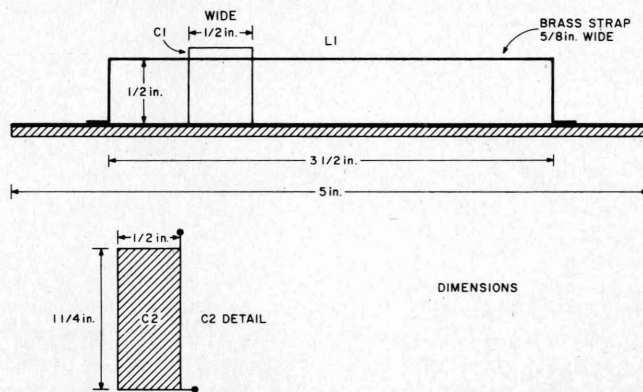


Fig. 6(b).

Fig. 7. Crystal exciter, 1220 MHz. Note that BB is connected to +12 V along with L5, as in Figs. 6(a), 6(b), and 6(c). (See lower left: L5 detail — 1220 doubler.) C out is placed 7/8" from end of L5 and is variable. The col. lead to L5 is only 1/4" long. CE is also similar to the emitter cap of Figs. 6(a), 6(b), and 6(c). BB is 5/2" long, 1 1/2" wide, and insulated from ground. L1 = 4 turns no. 16 bare, 3/4" long; tap at 1 1/2 turns from cold end, 5/8 o.d. L2 = 4 turns, 7/16 o.d., output tap at 1/2 turn from cold end. L3 = 3 turns; tap at 1/2 turn. L4 = brass or copper strap, 1/2" wide, 1-7/8" long, output tap at 1/2" from cold end, spaced 3/8" from baseband.

on the diode wavemeter. Fasten a ruler to the baseboard (be sure it has a centimeter scale), and, as you slide the short along the lecher line, take an average of several peaks. These peaks or dips are found at the half-wave points, so multiply by two for the wavelength. Easy-to-remember centimeter-to-MHz conversion points are 30 cm = 1000 MHz, 20 cm = 1500 MHz, 15 cm = 2000 MHz, and 10 cm = 3000 MHz.

I had to start using strap-line circuitry here at 610.8 MHz, as shown in Fig. 7. Around 400 to 500 MHz is where you have to give up on coils. From then on up, shape

begins to play an ever-increasing role in frequency determination.

The last stage of the exciter is where things get a little tricky if you are not used to half-wave double-short circuitry. However, once you get it running, you will find it tunes very sharply and provides you with good output on 1221 MHz. Note the small-value trimmers found between the base and ground of the last three doublers. These are quite important, as you will find, serving to match the base input impedance to the previous collector circuit, along with the tap-down on its induc-

tance. There is also a parallel capacitor, CP, across L5, the 1200 MHz tuned circuit. This is the same as on the lower frequencies, only a great deal smaller. A certain amount of parallel C combined with the L gives the best Q for a given frequency. A few minutes of trial, varying the C against the length of L5, will show you that that principle still holds true even on 1220 MHz.

The collector connection is shown tapped down on L5, also as usual. Note also the use of +12 volts on the ground plane of L5. This method avoids the use of "zero inductance" capacitors at the ends of L5. Such capacitors do not exist in

stores, although certain expensive coaxial types do attain quite low values of L in the "leads."

The emitter bypass, CE, uses the brass-plate method to achieve the desired low inductance for use at 1220 MHz. The oscillator, Figs. 6(a) and 6(b), shows more detail of this type of flat-plate cap. In those figures, it is labeled C2 and is used as the base bypass. Note that to make an oscillator, you "lift" the emitter off ground and ground the base (to rf, that is). To make an amplifier or doubler, you ground the emitter and use the off-ground base as the open input.

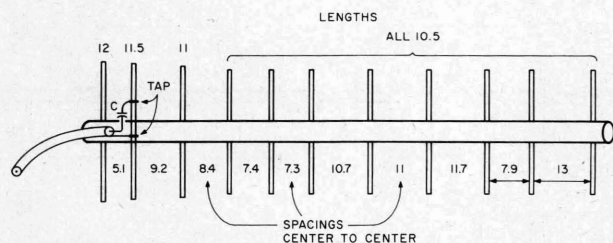


Fig. 8. 11-element beam, 1220 MHz. (All measurements are in cm.) Boom = PVC grey plastic tubing, 7/8 o.d. Elements = bare copper, no. 12. C = $\frac{1}{2}$ to 1 pF. Tap, center to center = 19 mm. Spacing, C, and leads to element = approximately 2 mm. Elements fastened with push-fit and white glue.

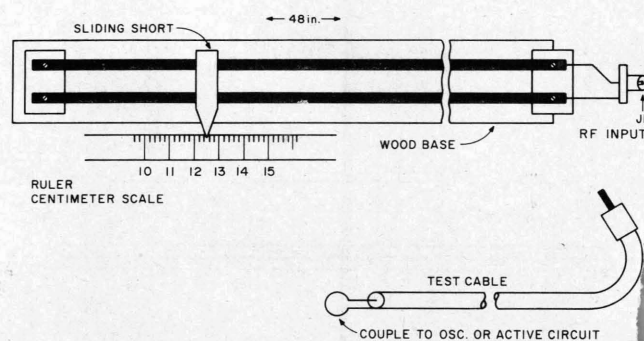


Fig. 9. Lecher line.

My usual method of tuning up a rig like this is to adjust each stage for maximum gain and output, making very sure of the frequency, before going to the next stage.

Low-Cost Yagi for 1220 MHz

The tune-up of a long Yagi for 1220 follows about the same procedure as on the lower frequencies, using a source (which in this case is the exciter of Fig. 7), a receiving beam, a tuned diode

detector as in Fig. 3(a), a microammeter, and various tabletop beam-element-stand pieces of hardware (made out of wood this time). You won't need the latter, as I have already done this work for you. Just follow the dimensions given in Fig. 8, and it will work fine. The only item that may take a little experimenting with is the gamma match. The small-value C, the tap on the radiator, the spacing from the radiator of the leads, and the

"dress" of the coax cable as it attaches to the radiator and the C are somewhat critical. Follow Fig. 8 as closely as you can, and experiment a little, and you should be okay. Just remember that most worthwhile things take a little time.

The beam itself, as shown in Fig. 8, is easy enough except for that gamma match.

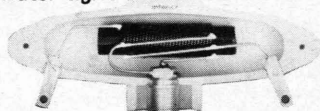
Be sure to keep your eye on the receiver meter, with

the receiving beam several feet away. When finished, mine worked across 10 feet of room, tabletop to tabletop, with the diode receiver. When a sensitive receiver is used, you can start hill-topping and get a first approximation of the range in kilometers for an "under 100-milliwatt" rig.

Then you can go on up from there, with more power, bigger beams, better receivers, and so forth, as your pocket-book and skill allow. ■

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A Brass Horn For X-Band

—simple 10.5 GHz antenna

Articles I've written about antennas for an X-band (10.5 GHz) transceiver and the "Smokey Detector"¹³ have brought many questions to my desk. One question in particular, asking where the horn used in the tune-up procedure for the Smokey Detector was procured, was the impetus for this construction article. Another question asked about the use of rectangularly-shaped waveguide anten-

nas as signal sources for equipment using circular polarized horns.

When microwave communications in the amateur segment of X band became my prime interest, equipment — mostly waveguide components — was easily found on the surplus market. Signal generators consisting of klystrons and power supplies were readily available after World War II, as were all of the nice goodies like slotted

lines, detectors, and precision attenuators. Many gain standard horns found their way to this market but were ignored, and the famous Polaplexor was used instead. That wonderful source of components has dried up, and so we are now forced to borrow from a friendly microwave man or do our own construction.

Horn antennas and all of the other equipment you will need to operate in this ex-

citing band, you can construct yourself. You don't need a machine shop and a lot of cash to buy the parts. Patience and careful use of simple hand tools provided the first experimenters with these items, and so they can provide for you. I made my own pyramidal horn, and here is how you can make a copy of it.

Before I get into the construction details, I will answer the second question mentioned above. Certainly there will be a signal loss over the path between a circular polarized horn and a rectangular or linear transmitter. The Smokey Detector tune-up procedure used these dissimilar polarizations because path loss was not a problem over the short distance used. This practice is not encouraged for regular communication unless you feel that long CQs will afford you the answers from the midnight DX on this band.

A search through known references related to microwave antennas reveals that a pyramidal horn will provide a pattern in the E plane which is nearly circular. This is the reason

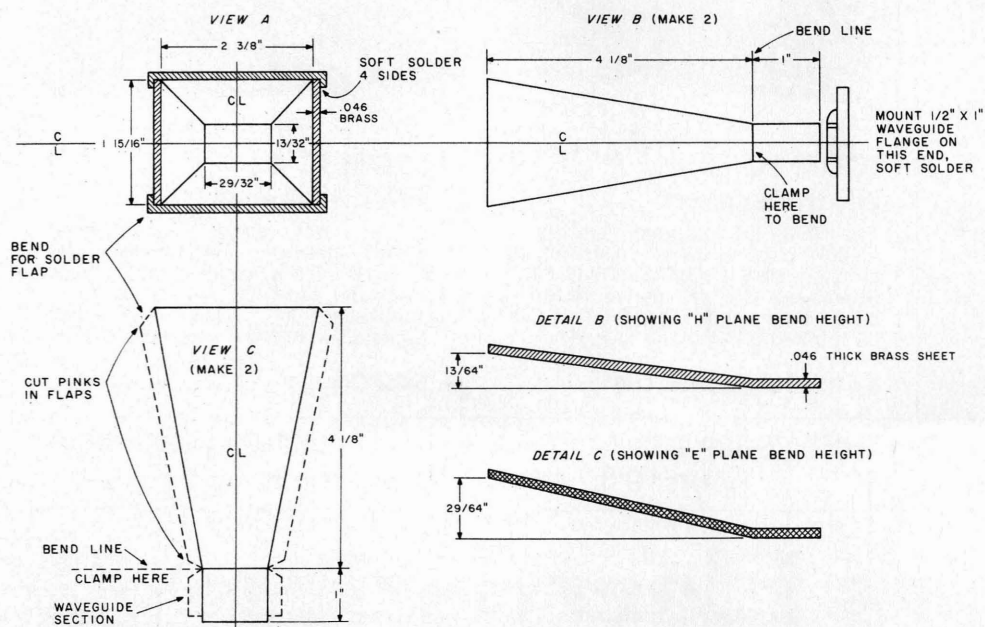
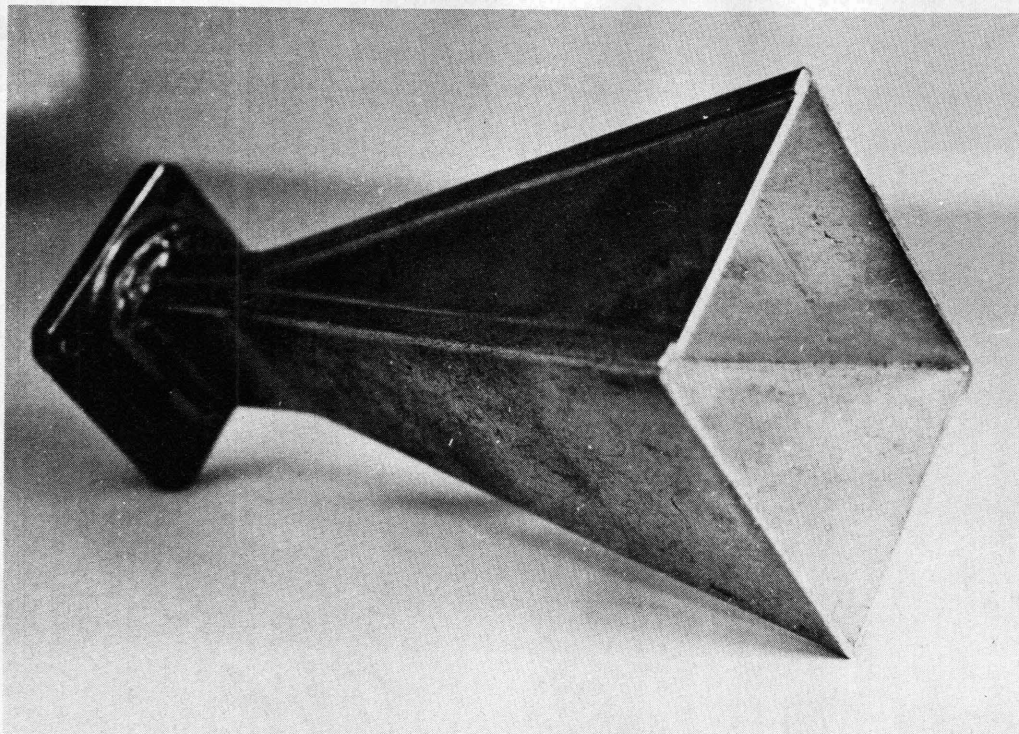


Fig. 1.



Pyramidal horn. This oblique view of the horn shows the assembly and the addition of a choke flange.

other section of the E-plane wall. Use a C-clamp to hold the whole assembly together by clamping across the E-plane section of the waveguide. Small pieces of thin aluminum sheet can be bent to form cross braces at the inside of the mouth of the horn and will aid in the soldering process. (A piece of wood shaped to fit in position will also do the trick.)

Clamp the outside edges near the end of the flare where the flaps meet the matching side. If you do not provide a firm clamp at these points, when the metal is heated for soldering, it will flex. If you are satisfied that the horn is clamped well enough, you are ready to solder it.

The easiest way to solder the assembled horn is to stand it up on the face of the open-flared end on an electric hot plate. It will take much longer to do the job than with a blowtorch or propane soldering iron, but the heat will be more uniform and the control over the way the solder will flow is better.

When the horn has been heated enough, which will be

indicated by the color of the metal where the soldering paste makes contact, place a piece of wire solder on one seam in a stroking action. The solder will run into the seam and down to the open flare. Make sure that just enough solder has filled up each seam to join electrically the seam edges. Use the solder sparingly; too much will cause lumps which will have to be removed. Control the heat by lowering the stove temperature, but keep it just hot enough so that the solder will run.

When this task is complete, turn off the heat and let the whole assembly cool off. Do not lift it off the heater and cool it quickly under water; flexure of the seams can open them.

Remove all jigs and clamps. Carefully inspect the seams to see that there are no gaps in the solder. If there are, the assembly will have to be reheated and the soldering trick repeated. If the soldering procedure has provided a solid connection between all parts of the seams and there are no gaps, fit the flange to the open waveguide end and

solder it in place squarely. Be careful not to disturb the previously soldered seams.

The completed horn should be washed in very hot water to remove all traces of the acid solder paste. It then may be painted with Krylon paint to keep the brass from corroding. Do not paint the flange face.

Testing the horn to determine its gain and field pattern requires the use of a signal generator, an attenuator, and two other similar radiating devices.

Connect one of the radiators to the signal source output and the second to the attenuator and a suitable receiving indicator. The separation between the two test setups at this frequency must be 10 feet. The path between them should be free from obstructions and reflections. A pair of ladders six feet high will serve, if two flat surfaces are placed on the top of them. Fasten the radiating horn firmly directed toward the second platform where the second setup must be arranged so that it can be rotated horizontally about the axis of the transmitting

horn. A scale laid out on a piece of polar graph paper will assist in locating the half-power points of the horn.

Turn on the equipment. Check that the receiving setup is performing and that sufficient signal is detected when the attenuator is set to 10 dB. Now find the true axis, and make a mark on the plot sheet. Turn the attenuator to the 10 dB position, and note the indicator level. Now rotate the receiving device to approximately 10 degrees off the true axis and note the level. Readjust the attenuator to bring up the level to equal what it was when the device was on the true axis, and note the attenuator difference. You are looking for the level which corresponds to the 3 dB or half-power points. It will probably take several tries using this technique to locate this position. When it is located, move in the opposite direction from the true axis, and locate the opposite half-power point. Mark this on the polar plot sheet.

When the measurements of the two similar radiating devices have been completed, substitute your new horn for the receiving device. Make sure that it is receiving in the correct plane. When it is in place, direct it toward the transmitting radiator centering the true axis through the center of the horn. You will see that the receiving indicator is off scale, or at least reading upscale, showing that the horn has gain. Reduce the indicator reading by adjusting the attenuator so that the indicator reads the same level as the original measurement. Note the attenuator reading and be sure to note which plane the measurement was taken in. This difference is roughly the gain over the original radiator. If the comparison radiator's gain is known, then you may calculate accurately the gain which will be near 13 dB power gain.

Repeat the same procedure in the other plane.

it was chosen over the more easily constructed sectoral horn. These references are listed at the conclusion of this article and will also assist you in confirming the pattern measurement.

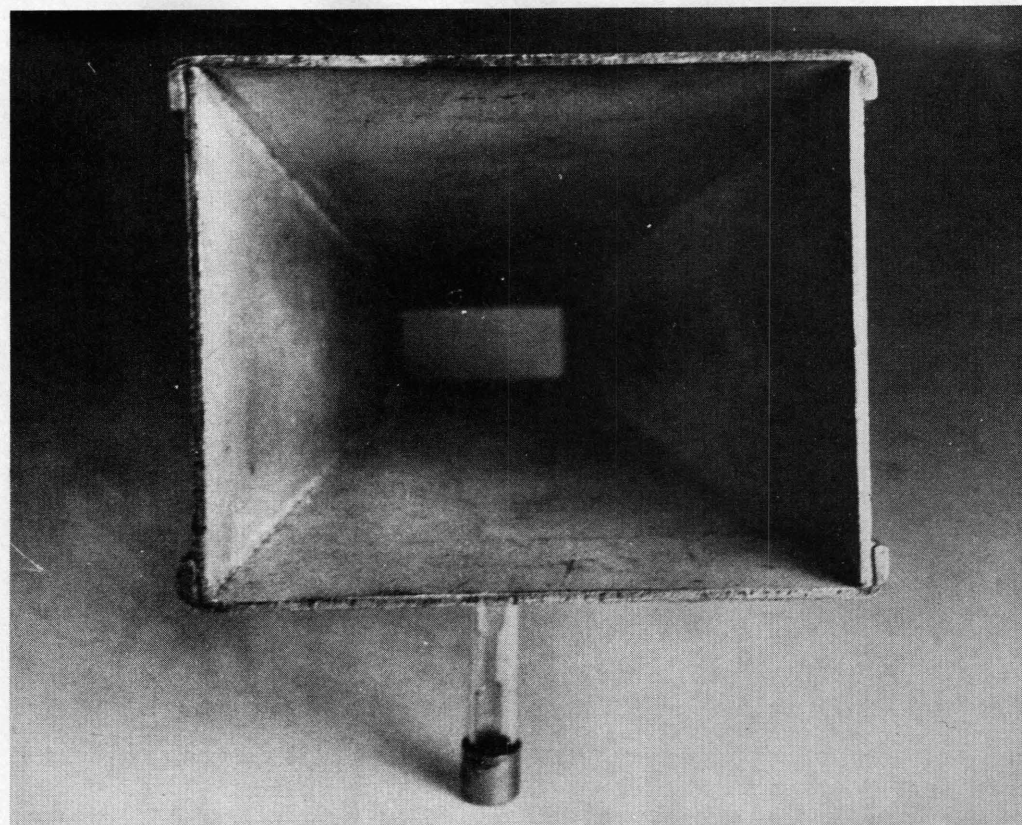
The construction materials needed are .046"-thick brass sheet, soft solder, and one 1/2 x 1-inch waveguide flange (UG-135/U). The inside dimensions of the flange are exactly 1/2 x 1 inch. (A 1.25" departure from the dimensions shown will result in a lot of file waving at the corners of the waveguide end of the horn.) The dimensions of the waveguide end are exactly those of a section of WR-90 small X-band waveguide.

A piece of hardwood two inches long, cut to a rectangular shape which will slide fit into a section of small X-band waveguide, will be required. This piece will serve as a clamping support when the final assembly and soldering takes place. Be sure that the piece is true over the full length and that the corners are smooth, or it will stick when you try to remove it. Starting dimensions for this jig are those shown at the throat of the horn in View A of Fig. 1.

The sheet brass for the sidewalls of the horn must be flat. If it is not, you will find that the horn will be very hard to assemble, because all edge surfaces must touch during the final assembly. Be sure that this material is clean and shiny. A small ball of steel wool can be used on the edge surfaces to insure that it is. When you attempt the final assembly, you will appreciate this extra effort.

Lay out the plates for all sides on the clean sheet brass. Scribe the positions of the bend lines clearly. If possible, use DyChem, a blue dye, on the brass so that the lines can be seen easily. Remember to wash off the dye with alcohol before attempting to solder. Allow twice the thickness of material for each bend

the dimensions given in View C (Fig. 1) do not in-



Mouth view of an X-band horn. This shows how the folds are made. Note the smooth inside surface.

clude the material needed for a solder flap, so allow at least 3/8" more material for this purpose. Also, cut 45 degree "pinks" in the flaps at the junction of the horn flare and the beginning of the 1-inch waveguide section.

The use of a sheet metal shear and brake will simplify the next few operations, but, if these tools are not available, two pieces of 1" angle iron and a large vise will serve to do the job. The cutting should be done with a fine-bladed hacksaw. The edges of the cuts can then be smoothed with a file.

Cut two pieces, as shown in View B (Fig. 1), for the H-plane sides of the horn. Place one of these pieces on a flat surface, and clamp it securely to the surface with a C-clamp located at the intersection of the waveguide section and the end of the flare. Place a thick piece of metal under the C-clamp foot on the bending line, as indicated in View B. Now, using another larger piece of metal or a large putty knife, lift up

sharply at the mouth end of this section to produce a bend at the clamped end. The height of the bend must be 13/64". The metal is springy and will take some pushing and pulling to achieve the correct height. Care must be exercised during this process so that you don't deform the walls with bends or deep scratches.

Follow the same procedure for the three remaining sections of brass.

When all of the forming bends have been completed, the solder flap bends are next. This job is done in two moves. Place the aforementioned sections of angle iron in the vise jaws. Then station the waveguide end of one E-plane plate between the angle irons so that the bend line of the flap is in position for a bend to 90 degrees.

To make this bend, allowances have been made for the thickness of the material, if you have followed the layout instructions. If not, the width dimension will come out wrong. Check now, and save

some grief. When you have determined that the measurements are the proper dimensions, make these bends so that they are square, right-angle bends. If the bend is too shallow, it will make the horn flare wider in one plane, and, of course, if it is too great a bend, the opposite effect will be the result. A 90-degree bend, therefore, is the desired angle.

After completing the waveguide bends just described, complete the horn flare flap bends on the E-plane section. Do both pieces at this time.

Check that the bending you've done has resulted in square and true wall sections, and make sure all the edges are clean and ready for solder. Apply a moderate amount of acid soldering paste to the inside of the flap bends. Now place the previously-prepared wood jig inside of the E-plane waveguide bends of one section. Slip the two H-plane wall sections into place on each side of the jig block. The last piece is the

ways making sure that the antennas are in the same plane. Cross polarization will be easily detected by the very weak signal received.

This completes the construction of the horn and its measurements. You should have an antenna with a beam-width of nearly ten degrees to the half-power points in the E-plane and a little wider in the H-plane.

The references which appear at the end of the article are required reading if you

are going to attempt this project.

I must sound a note of caution regarding a problem which may be encountered when power exceeds the milliwatt region. It is a well-known fact that radiation from antennas or waveguides which produce an illumination over human tissues in excess of 10 milliwatts per square centimeter can cause serious damage to exposed tissues. Persons who work on military radar can appraise

you of this danger and the many lectures they receive about the subject. The most important warning I remember, which I received during my training, was "Don't look into the antenna or into a waveguide." Cataracts on your eyes may be the result. So be careful; don't look into horns or waveguides or, for that matter, any of the UHF antennas you use. Some excellent reading on the subject is listed in the references.

I hope to hear you on

10.445 GHz. I keep a sked with WA1IKR at 0010 UCT on Thursdays to beat the QRM problems. See you there. ■

References

1. Silver, *Radiation Lab Series*, Vol. 12, Sections 15-9.
2. Krauss W8JK, *Antennas*, Chapter 13, Section 13-6.
3. Olberg W1SNN, "Mobile Smokey Detector," 73, Holiday, 1976.
4. Brodeur, "A Reporter At Large — Microwaves," December 13 and 20, 1976, *New Yorker Magazine*.

Social Events

om page 149

c/o Sue Hagedon WB8GWO, 1340 Brainard Woods Drive, Dayton OH 45459.

TUCSON AZ APR 28-30

The Tucson Hamfest will be held on April 28-30, at the Ramada Inn (just off north I-10). It will feature technical sessions with demonstrations, microprocessors, solar posers, QRP, fast/slow scan, RTTY, remote base, etc. There will be prizes, ladies' programs, a banquet, exhibits, and a swap meet. It is sponsored by the Old Pueblo Radio Club. For information, write: OPRC, 1361 E. Edlin, Tucson AZ 85711.

SPOKANE WA APR 29

SWAP-FEST '78 will be held all day on Saturday, April 29, at the Spokane Interstate Fairgrounds. Flea market, mini-auctions throughout the day, contests, family picnic, major evening auction, some most-unusual radio exhibits, valuable prizes. Sponsored by the Inland Empire area amateur clubs. Talk-in on any area repeater. Write: SWAP-FEST '78, PO Box 3606, Spokane WA 99220.

MEADVILLE PA MAY 6

The 4th Annual Northwestern Pennsylvania Hamfest will be held on May 6th at the Crawford County Fairgrounds, Meadville PA. Gates open at 8:00. \$2 prize ticket required. Admission — \$1 to display. Children free. Hourly door prizes; refreshments; commercial displays. Come. Indoors if rain. Talk-in on 144 and 52. Details: CARs, PO Box 16335, Meadville PA 16335.

LAS VEGAS NV MAY 12-14

3rd Annual West Coast VHF will be held at the Star-

dust Hotel, Las Vegas Strip at Convention Center Drive.

Conference highlights: technical program arranged by the San Bernardino Microwave Society, hospitality room, informal technical and operating sessions, noise figure measurements contest, antenna gain measurements contest, prize drawing, 24-hour adult entertainment! World-famous resort hotel with all facilities. Look for the Stardust sign east of I-15. Take the Sahara Ave. or Dunes-Flamingo exit. Advance registration fee is \$4.00 per person (\$5.00 at the door). Make checks payable to: West Coast VHF Conference, 510 South Rose St., Las Vegas NV 89106.

DEERFIELD NH MAY 13

The Hosstraders net will hold its fifth annual tailgate swapfest Saturday, May 13, at the Deerfield, New Hampshire, fairgrounds (covered building in case of rain). Admission is one dollar; no commission or percentage. Commercial dealers are welcome at the same rate. Excess revenues benefit Boston Burns Unit of the Shriners' Hospital for Crippled Children. Last year we donated \$430.80. Talk-in on .52, 146.40-147.00, 3940 kHz. If you have questions, send SASE to Joe Demaso K1ROG, Star Rt., Box 56, Bucksport, ME 04416 or Norm Blake WA1IVB, P.O. Box 32, Cornish, ME 04020 or check the Hosstraders net on Sundays at 4 pm on 3940 kHz.

WEST LIBERTY OH MAY 14

The Champaign Logan Amateur Radio Club, Inc., will hold its annual hamfest on Sunday, May 14, 1978, at the West Liberty Lions Park, West Liberty, Ohio. Free admission; trunk sales; tables are \$1.00. Door prizes. Talk-in on 146.52.

WARMINSTER PA MAY 14

The Warminster Amateur Radio

Club's fourth annual "HAMMART," flea market, and auction will be held on Sunday, May 14, from 9 am to 4 pm at William Tennent Senior High School, Street Road (Route 132), 2 miles east of York Road (Route 263), Warminster, Bucks County PA. Registration is \$1.00, tailgating \$2.00 additional. No indoor selling; bring your own tables. Talk-in on 146.16-76 and 146.52. For further information, write: Horace Carter K3KT, 38 Hickory Lane, Doylestown PA 18901 or call (215)-345-6816.

WABASH IN MAY 21

The Wabash County Amateur Radio Club's 10th annual hamfest will be held on Sunday, May 21, 1978, rain or shine, at the Wabash County 4-H fairgrounds in Wabash. Large flea market (no table or setup charge), technical forums, bingo, free parking, and lots of good food at reasonable prices. Advance admission is \$2.00; \$2.50 at the gate. Children under 12 free. Write Dave Nagel WD9BDZ, 555 Valley Brook Lane, Wabash, IN 46992.

COLUMBIA SC MAY 21-22

The Carolina Repeater Society is sponsoring the Columbia Hamfest on Saturday and Sunday, May 21 and 22, from 9 am to 5 pm at the Jamil Shrine Temple located 1 mile west of I-20 on

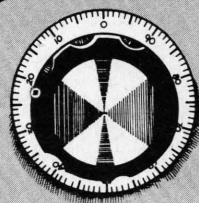
I-26. Large air-conditioned building with plenty of free on-site parking, a flea market, dealers, and activities. Talk-in on 34/94. Combined admission and drawing tickets are available for \$3.00 in advance or \$3.50 at the door. Contact Larry Johnson WA4VOJ, 1520 Atlantic Dr., Columbia SC 29210, or phone (803)-772-7984 or (803)-788-1308.

ERLANGER KY MAY 28

The Kentucky Ham-O-Rama will be held on Sunday, May 28 (Memorial Day weekend), at Erlanger Lions Club Park, Erlanger, Kentucky. It's 7 minutes south of Cincinnati, Ohio, 1 mile off I-75 south, the Donaldson Road exit. Talk-in on 146.19-79 repeater, 52-52 simplex. There will be prizes, exhibits, and a flea market. For information: NKARC, Box 31, Ft. Mitchell KY 41017, or phone (606)-331-4922.

FT TUTHILL AZ JULY 28-30

The Amateur Radio Council of Arizona will present the annual Ft. Tuthill Hamfest on July 28, 29, and 30th, 1978. Come on out in the cool pine country of Arizona, and join our western barbeque, prize drawings, and tech sessions. For further details or pre-registration forms, contact PO Box 11642, Phoenix AZ 85061.



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Close Encounters

—the eyes of Texans are upon them

Something strange was moving up there! Across the glittering star fields of a moonless Texas night it crept, a small orange light, pulsating slightly and growing brighter. Abruptly, it changed direction. Reddish now, it proceeded at right angles to its former course, away from the smudge of light on the horizon that marked a distant city.

A flight controller hunched intently over his radarscope. Its eerie glow illuminated an expression of amazed disbelief. A silent whistle escaped from his pursed lips. An 80° turn at 16,000 mph and out of range already? Involuntarily, his throat muscles tensed to speak to the pilot of the only plane on the scope, then relaxed. Who would believe him? Probably an equipment malfunction, he thought. Yet stories told by old-timers, stories at which he had scoffed, began to filter into his mind.

Much lower now, the object skimmed slowly over an area of rough terrain. A lone car probed the dark

county road with high beams. Nearing the crest of a hill, it switched to low as a glare showed someone was coming. The beer net on 34/94 was pleasant company. Suddenly there was only dead silence. Worse yet, the engine and headlights had quit at the same moment!

Too busy braking to question the source, the driver was thankful for the light as he brought the car to a stop on the berm. But now the approaching blaze looked like a jet-propelled magnesium flare. Just as it seemed that it must smash right into the car, it was up and over and off into the sky behind. And a ham sat quietly, shaking for five minutes before realizing that the engine was running, the headlights were on, and the repeater was chattering away as though nothing had happened. "What was *that*?" was still his only thought.

At that moment, not far away, as it had all day, every day for months, a unique laboratory waited to answer that question. Near the very limit of their sensitivity, recording instruments deviated slightly

from the norms of their tireless monitoring. Inside a low building, pale by starlight against the dark hillside, electrons surged through microcircuitry. A minicomputer swiftly executed its intricate series of commands. An alarm shrilled, alerting duty personnel. Quickly all posts were manned, and the sophisticated technology of the only known scientific facility in the world dedicated solely to UFO research was ready for what might come.

Still adjusting headsets, observers manning three phototheodolites at widely separated locations on the 400-acre site scanned the stars for one that moved, waiting for instructions. They were not long in coming. "Magnetic anomaly, 270 degrees, increasing in intensity. Stand by." Inside the laboratory, the director studied the endless white tongue of paper extruding slowly from the chart recorder. Eight fine lines were being penned on it, measuring the output of various sensors. Periodic blips indicated time signals being received on 60 kHz from WWVB. Two of the

channels were now showing deviation well above their baselines. Attention shifted expectantly to the color video terminal.

From high atop a tower rising into the darkness above a nearby building, powerful radar pulses were sweeping a 12-mile radius. For several rotations there was no unusual return. Then, "Radar lock-on!" As coordinates of the UFO were relayed to the field observers, excited cries doubled in the headsets. "Got it! Orange lenticular object, moving in fast."

All three phototheodolites were now tracking the object. Each operator concentrated on keeping the image of the UFO centered on an illuminated spot in his aiming scope, while shaft encoders on the pan-and-tilt heads of the telescopic cameras were feeding coordinates into the computer. At the same time photographic evidence was being collected, data sampled from each of the three locations every few seconds was being processed into a video display. The UFO's path was seen superimposed over an im-

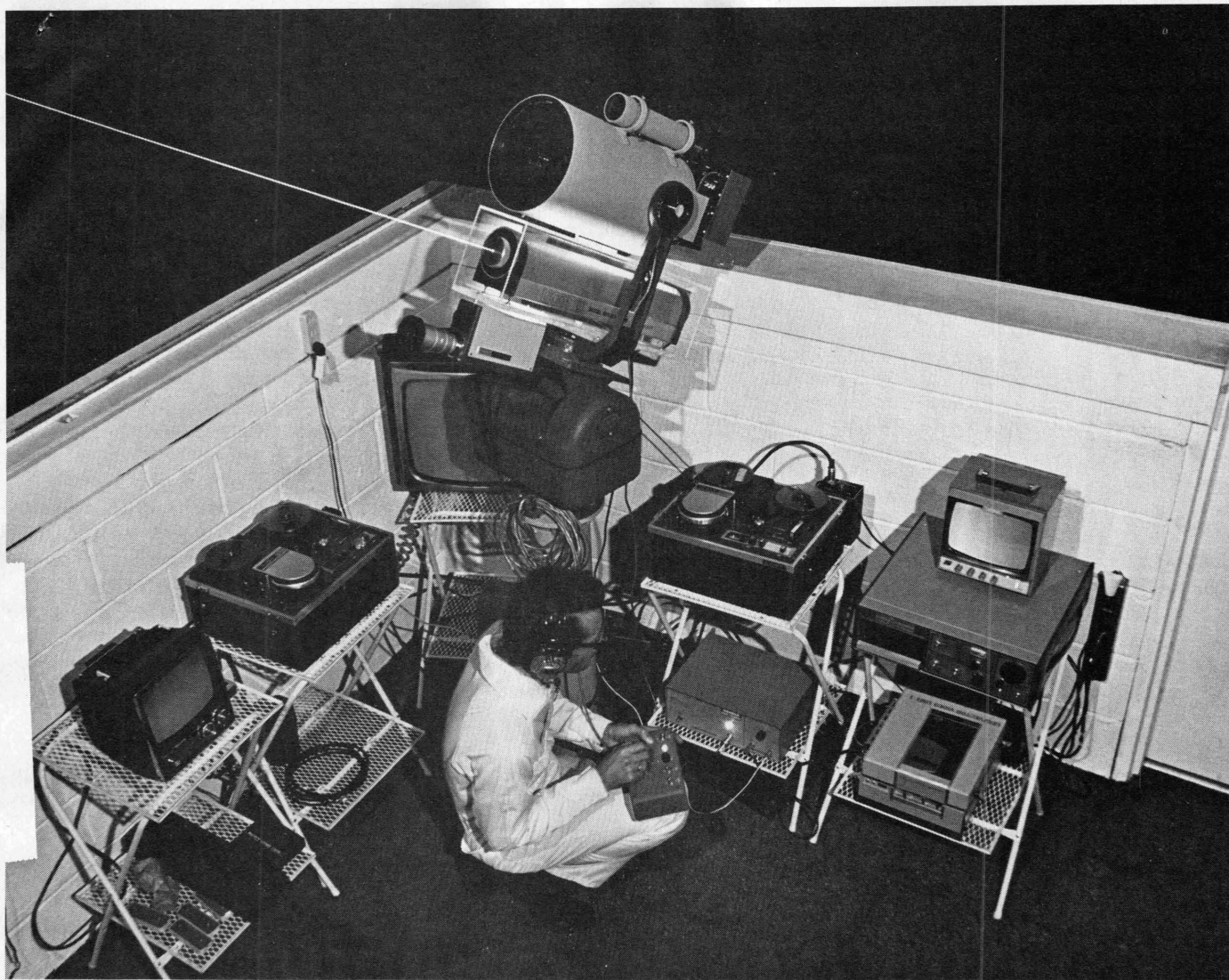


Photo A. Laser beam made visible by water vapor refraction pierces the night from the UFO light-pulse experiment apparatus at the Laboratory for Instrumented UFO Research near Austin, Texas. Capable of transmitting up to 2.5 million bits of response-test data per second, this red-light laser device can also be used to measure UFO distance and test the hypothesis that light beams may be bent in the vicinity of some UFOs.

age of the area beneath it. Actual distance readings were being printed out for permanent record.

Busy as it was, though, the computer was also performing a number of other vital functions. As the vidicon operator focused on the approaching UFO, it measured the arc subtended by the image and computed the size of the object. It also computed the visibility radius of the object, and retrieved the names and phone numbers of ARGUS volunteers who should be able to see it. Several telephone lines were being pulsed with the dual tones so familiar to

autopatch users, and sleepy voices began answering phones shrilling on bedside stands miles away. As each answered, the name and phone number was printed out and the volunteer heard, "This is an Operation ARGUS alert! Please do as you were instructed."

Suddenly wide awake, the observers hastily pulled on clothes, jammed feet into shoes, and grabbed binoculars and cameras on the run. This night they were not to be disappointed. Here was UFO event-sharing on a silver platter, in contrast to the ordeals suffered by

our friends in "Close Encounters" on the silver screen.

Unlike the movie, the scenario we have imagined is hardly fantasy. This laboratory actually exists. At this very moment, whatever the time, its equipment is scanning the sky, waiting for the real thing to happen. This is where history may be made—Project Starlight International, or PSI.

In the rattlesnake-infested hill country northwest of Austin, Texas, accessible only by four-wheel drive, lies the 400-acre site of the Laboratory for Instrumented UFO Research,

a facility unique in the world. At this remote location, field research is conducted for Project Starlight International, a research division of the Association for the Understanding of Man, which is a nonprofit educational organization based in Austin. PSI's purpose? To document scientifically and irrefutably the existence of UFOs. Ray Stanford, founder and managing director, is an acknowledged expert in the field of UFO research. Author of *Socorro "Saucer" in a Pentagon Pantry*, he conducted a fascinating and well-documented investigation of the Socorro,

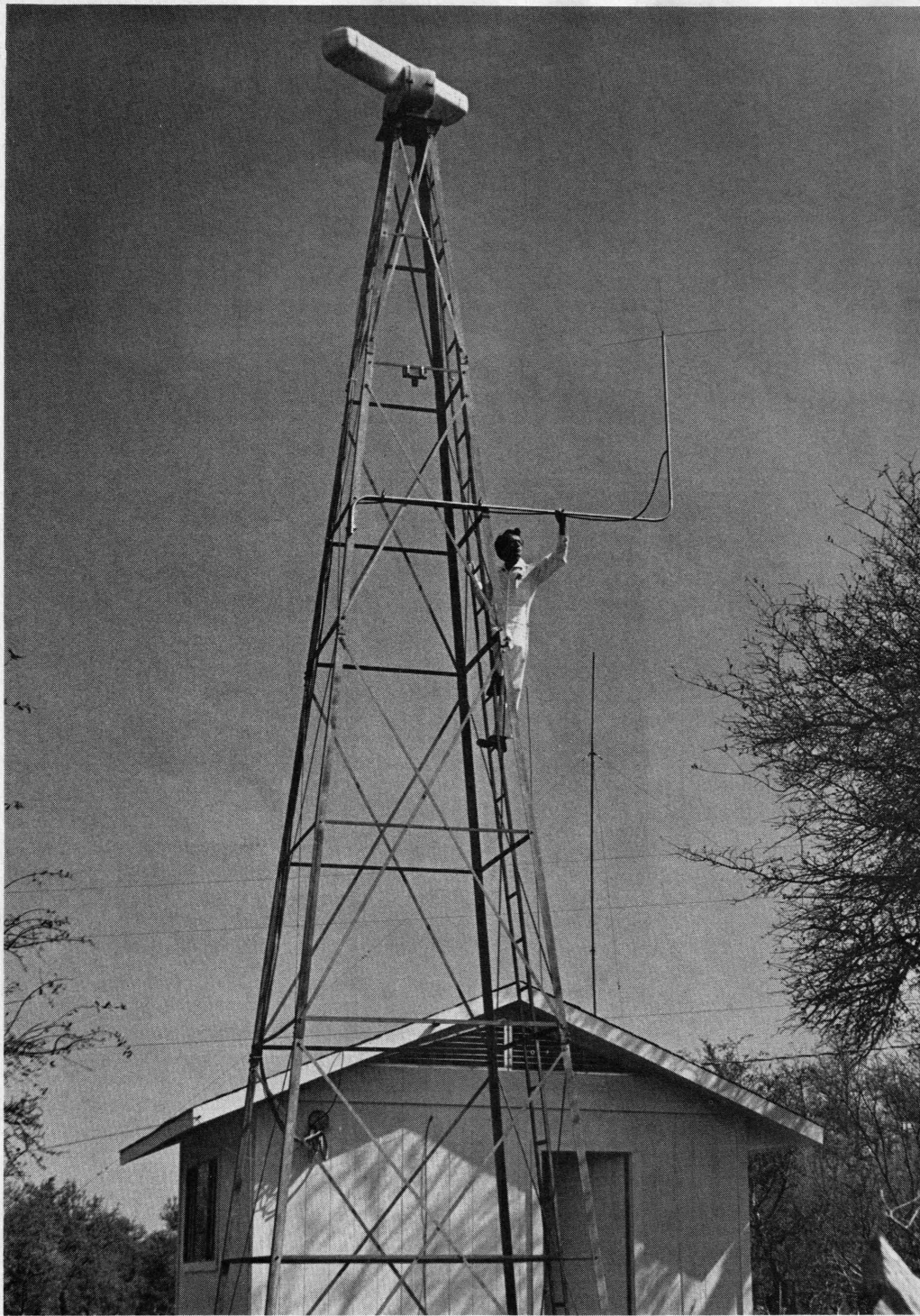


Photo B. PSI's Operation ARGUS radar and radio-frequency monitoring laboratory. With a radius of 12 miles, this radar unit will provide UFO distance data for Operation ARGUS, a computer-centered tracking system of highly sophisticated design covering a range of 472 square miles.

New Mexico, landing of April 24, 1964. According to Stanford, there is no known research facility in the world dedicated to UFO investigation which even approaches the sophistication and capability of PSI.

This high-powered re-

search effort is directed by a professional astronomer, Dr. Daniel H. Harris, Ph.D., from the University of Arizona. Dr. Harris, something of a modern pioneer, is the first scientist to accept a full-time paid position in UFO research. Right now, final touches are be-

ing completed on the most sophisticated of the equipment, and the laboratory will be fully operational. Much of the equipment is already scanning Texas skies twenty-four hours a day. And a most impressive array of scientific goodies it is indeed.

What are the prospects for irrefutably documenting a close encounter? Much better than you might think, as witness the photographs showing only one of several UFOs observed at the site. But wouldn't it be better to go to the UFOs rather than hope they appear at one location? Actually, that was the historical approach. During the green fireball episode in the late '40s and early '50s, teams of investigators for Project Twinkle rushed from one area to another where sightings were being reported. Invariably, they arrived too late to see anything. The UFOs, it seems, didn't wait around for them. PSI decided that it would be more productive to establish a permanent laboratory with sophisticated equipment and man it around the clock, seven days a week. The other option is still open, however. A vital core of instruments can be transported on short notice by four-wheel-drive van to any location where it might be needed.

Until now, most UFO research has been anecdotal. Witnesses of past events could be interviewed and second- or third-hand information could be correlated. Infrequently, a fortuitous amateur photograph, usually of very poor quality, might turn up. Or perhaps a bit of soil from a purported landing site could be secured for analysis. Immense effort went into analyzing and rehashing data of this kind, and there is a lot of it. UFOCAT, the computerized files associated with the Center for UFO Studies, now contains over 60,000 close encounters. And Ted Bloecher has indexed over 1500 close encounters of the third kind, in which contact with entities was reported.

However, there was no way to study UFOs directly and scientifically. Like the weather, lots of people talked about UFOs, but nobody did anything about them—except for the military, which was busy trying to shoot them down.

Scientific voices have cried in the wilderness almost from the beginning of the modern UFO era in World War II, urging serious investigation. Back in 1968, the House Committee on Science and Astronautics held a hearing on UFOs. Dr. Garry C. Henderson, then project leader on the lunar surface gravimeter/surveying system, proposed an implemented plan to acquire hard facts about the existence and nature of UFOs. He even detailed the instruments which should be used. And Carl Sagan, an astronomer who is as outspoken an advocate of the well-inhabited universe theory as he is a skeptic about UFOs, has said that anyone really interested in the supposed phenomenon should use high-quality instrumentation to probe its nature. Finally, someone is doing just that!

PSI is equipped to study a broad range of physical effects which might be associated with UFOs. Their objective is to gather a maximum range of hard data and to disseminate this information quickly to members of the scientific community. At a local level, larger numbers of people, probably including some hams, will be able to share in UFO events through Operation ARGUS.

The Greeks, as usual, had a word for it. Argus was a character in Greek mythology who had eyes all over his body to make him a good watchman. At the Laboratory for Instrumented UFO Research, ARGUS stands for Automated Ringup on Geo-

located UFO Sightings, and we have illustrated how it might work in practice. But there is a lot more to scientific UFOlogy than this.

UFOs have been reported to cause magnetic, radio-frequency, electrostatic, and gravitational effects, as well as temperature changes, barometric disturbances, and sounds. PSI's automatic recording equipment therefore includes three magnetometers and a gravimeter, as well as a microbarometer, an electrometer, and a sky camera activated by magnetometer deviations. An ambient microphone records voice input and audio effects, while a highly-directional microphone can handle distant sounds. The eight-channel, sensor-activated chart recorder displays low-frequency data up to 150 Hz correlated with universal time from WWVB. Radio-frequency scanners and recorders also incorporating UTC input cover the rest of the spectrum.

A computer-interfaced magnetometer system has been completed which will process field-effect data. Newly-designed sensors with 60-Hz filters respond up to 700 Hz and are oriented in three dimensions. Thus a three-dimensional video model of the magnetic field around a UFO can be displayed, showing each component in a different color. Pulsations or changes in light emitted by a UFO can be monitored by an electronic system utilizing solid-state sensors having a bandwidth of 10

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Hammond IN 46325

James Lorenzen, Director
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Tucson AZ 85712

Table 1. Major UFO research organizations.

MHz.

OZMA and CYCLOPS are strange-sounding names for serious projects funded by the U.S. government to search for intelligent life in space. Possible communications from selected stars have been monitored. SETI, Search for Intelligent Life, is an ongoing NASA project which is developing designs for a very large system of antennas and computers for the purpose of contacting extraterrestrial life. Since NASA scientists are convinced of the importance of such endeavors, PSI has not neglected this aspect of UFO research. Are UFO intelligences, if they exist, capable of or interested in exchanging intelligent communication? To answer this question, a

modulatable Liconix 605M helium-neon laser has been installed which can transmit voice, code, or television signals. Any modulated light response which a UFO might make to the laser signals can be detected as sound or as a TV image.

Radio transmissions other than noise have not been reported from UFOs. Disruption of radio transmission and reception, on the other hand, is frequently reported. This is why laser light rather than rf was chosen for a communication experiment. According to many reports, what appear to be coherent light beams of various colors have been projected from UFOs. And searchlight beams directed at UFOs have been seen to bend sharply, due perhaps

DAY	BAND	MHz	UTC	EST	NET CONTROL	QTH
Saturday	40	7.237	1200	0800	N1JS	MA
Saturday	75	3.975	1300	0900	WA9ARG	IL

Table 2. MUFON amateur radio SSB nets—weekly.

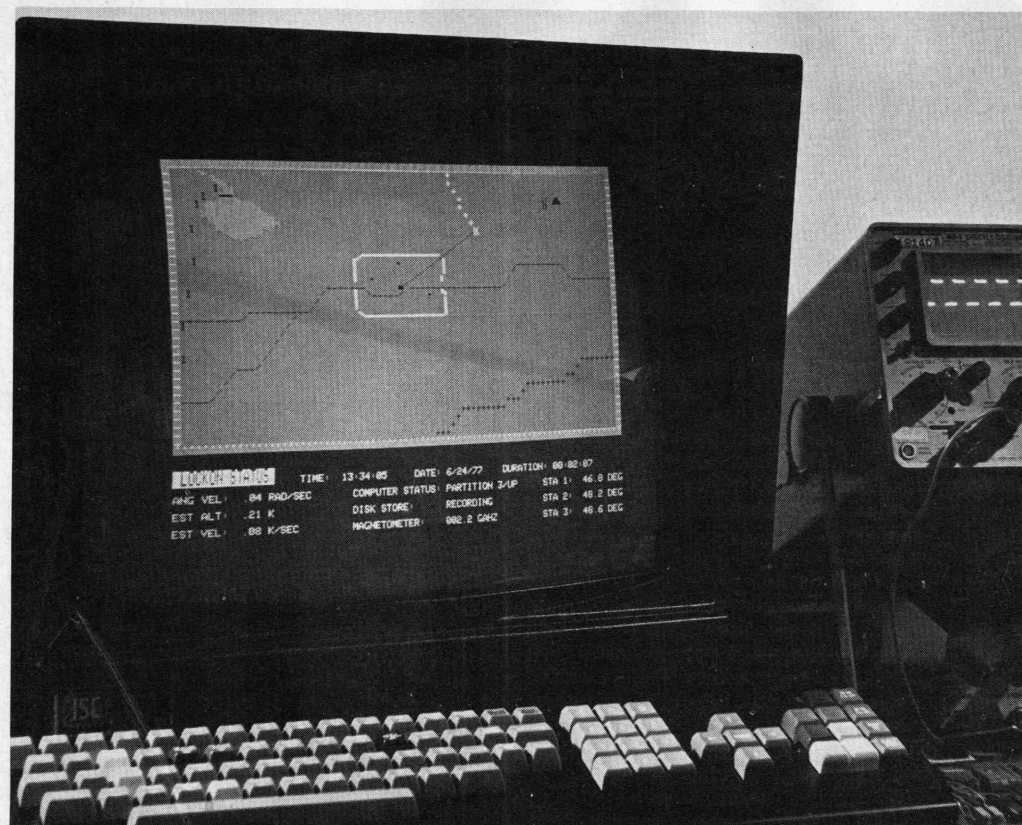


Photo C. An early prototype program display on Operation ARGUS's color video terminal, showing simulated UFO tracking over a computerized topographic map. Tracking and laboratory instrument data is automatically displayed below the map.

to some field effect or variations in atmospheric density. These are phenomena worthy of investigation for which the laser equipment could be used. In addition, the system can be adapted to determine the distance of an object with extreme accuracy using reflected laser light.

As you might expect, photographic documentation is an important aspect of the laboratory program. UFOs within range will find that they are captured on 35mm moving picture film. A Super-8mm sound movie camera with a 1-to-12 ratio zoom lens is also available. From various stations at the site, three automatically synchronized 35mm cameras, one of which is equipped with a diffraction grating for spectral studies, record any UFO event. High-resolution, close-up images of objects being tracked can be obtained

using Schmidt-Cassegrain telescopes of 2,110mm and 1,250mm focal lengths, as well as a 240mm telephoto lens on the 35mm movie camera.

UFOs have often been reported to investigate new or unusual light patterns on the ground. Some have responded to lights flashed or directed at them. For this reason, a light pattern response experiment has been devised, although it is rarely used. A hundred-foot circle consisting of ninety-one 150-Watt spotlights contains a single light in its center. Solid-state circuitry and a microprocessor make it possible to sequence the lights in any desired pattern, or even to mimic the light patterns of a UFO.

K12XBJ, the only known radar facility in the world dedicated exclusively to UFO research, was licensed by the FCC on June 8, 1977. Although it is

planned to install a more effective system for broad-range sky coverage when funds permit, the present Raytheon Model 1700 covers a 12-mile radius with 360-degree rotation. Operating on 9375 MHz, its 7.5 kW pulses can detect reflective objects up to 20 degrees above the horizon.

How big was the UFO? This easy-sounding question is one of the most difficult to answer accurately when a sighting has occurred. Was the object very large and far away, or was it small but close to the observer? Few people run around with optical range finders in their pockets, and it is rare that a UFO passes in front of some background object which can provide a distance reference. At PSI, however, Operation ARGUS can determine distance electronically by radar. Not all UFOs reflect radar signals, apparently,

but this poses no problem. Accurate horizontal and vertical coordinate data from shaft-encoders on optical tracking equipment can be triangulated by the computer to provide actual distance, horizontal distance, and altitude. If the area of an image can be measured, the size of the object can then be computed from the distance data.

When a UFO is being tracked, the ARGUS computer has been programmed to select from its memory of 472 square miles of terrain that sector of a full-color topographic map over which it determines the object to be passing. The path of the UFO then appears on the video display superimposed over the image of the terrain. Sequentially-tracked positions are indicated by successive letters or numbers. The entire episode, correlated against UTC, can be retrieved from computer memory for later study. Ground objects over which the UFO passed or hovered as well as as possible landing sites will thus be a matter of record. They can be examined for evidence later, if the UFO departs before a mobile unit can reach the site.

We've had a look at the GUS of Operation ARGUS, which is primarily technological. The AR, automated ringup, deals with people, for it is in this way that local volunteers can get involved. Ray Stanford terms this aspect of the operation "UFO event-sharing." Here is a concept of great potential to us as amateur radio operators, wherever we may live. As a movie, "Close Encounters" was great entertainment and could even be considered educational in some respects. But, fantasy aside, what is the actual status quo with regard to UFO knowledge at the

present time?

To be honest about it, there is a great diversity of opinion on the subject. UFOlogists, many of whom have been investigating the phenomenon for thirty years, present a spectrum of opinion. Some take the position that little or nothing is known concerning the true nature of the UFO. Official government interest vanished with the dissolution of Project Bluebook and the issue of the infamous "Condon Report," which as much as denied their existence. On the other hand, a number of authorities believe that the reason for governmental disinterest, including the recent refusal of NASA to reopen the field for investigation, is that they already know all about UFOs. In his book, *Situation Red: The UFO Siege*, Leonard Stringfield builds a strong circumstantial case that intact spacecraft have been recovered from crash sites, and that extraterrestrial humanoids have been autopsied. If so, it now appears unlikely that military authorities will voluntarily expose these facts to public view. However, a lawsuit filed by one UFO group against a government agency under the Freedom of Information Act could produce evidence of such concealment.

Between these viewpoints, one finds many theories about the nature of the UFO. Some UFOlogists believe that the phenomena may be psychic in nature. Others think UFOs are a mass neurosis, a psychological projection from the race mind. A few like the idea that they are a control mechanism, designed to influence human evolution in the manner we saw dramatized in the movie "2001." Most, however, believe the evidence points to hardware from

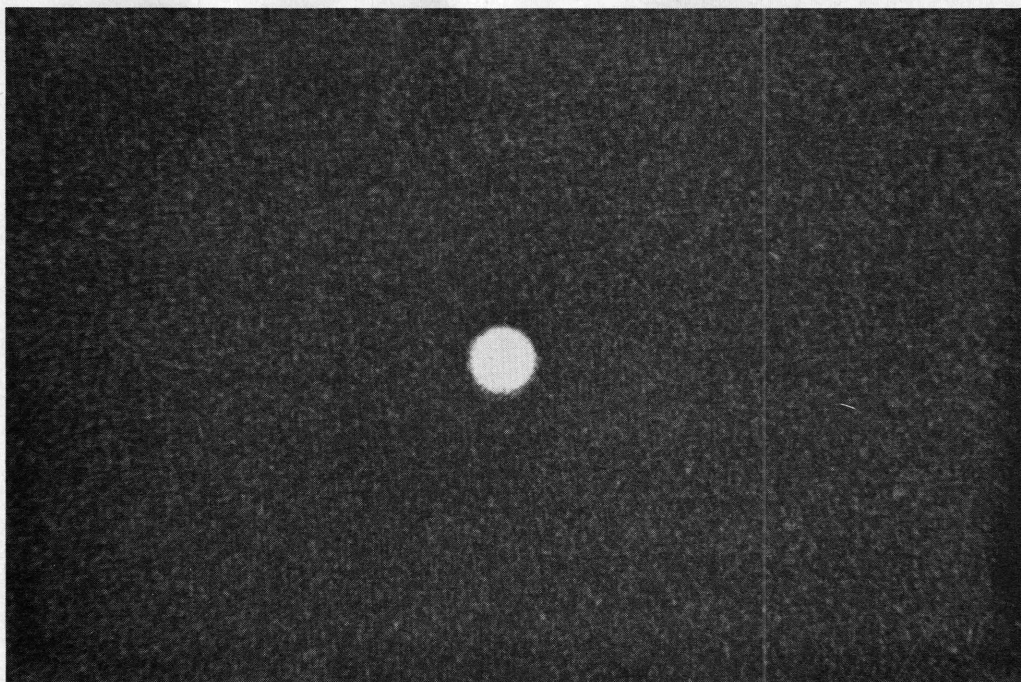


Photo D. Second-generation print of possible UFO which hovered for nearly 10 minutes beginning at 8:58 pm on December 10, 1975. Tri-X film, 5-second exposure with 300mm f/4 lens. Forty-eight photos were obtained during this event, which occurred prior to installation of PSI's more sophisticated equipment.

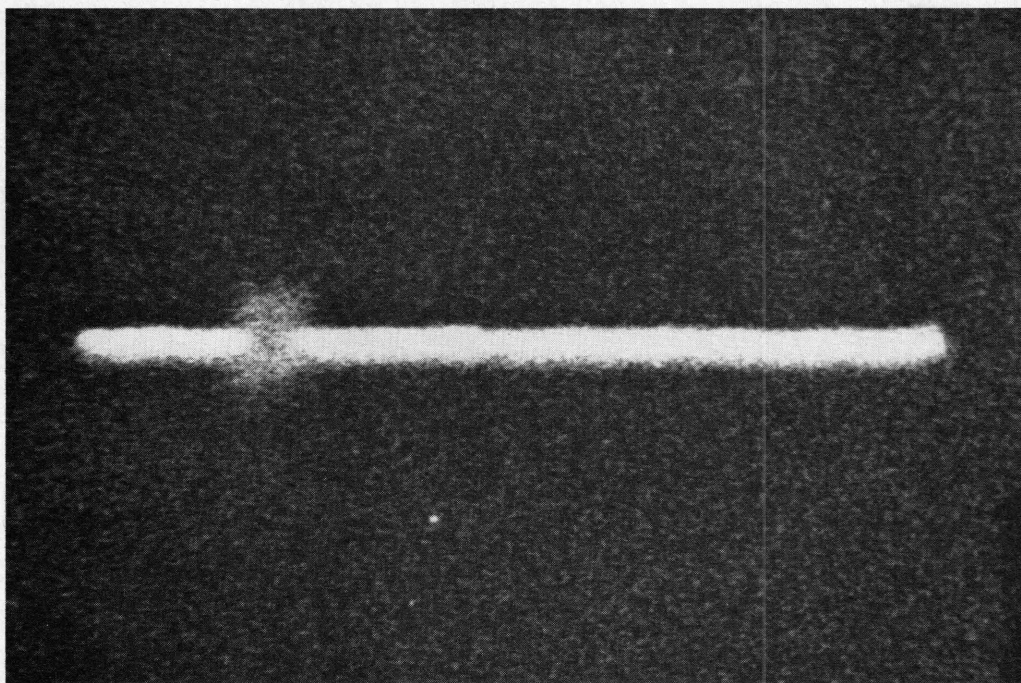


Photo E. Second-generation print of same object as Photo D moving off to left during 8-second exposure. Note strange burst-like effect not apparent to observers. Approximately 9:09 pm, Tri-X film with 300 mm f/4 lens. Typical of those taken by the PSI staff during the December 10, 1975, event, the photos are encouraging but not considered definitive concerning the nature of UFOs.

outer space, vehicles from some distant star system which operate through space/time in a manner we cannot yet comprehend. It is this hardware aspect of

UFOs which renders them susceptible to instrumental investigation. We may be on the way to answering what UFOs are, but the questions of where they

are from and why they are here will ultimately have to be answered as well.

Where, then, does all this leave us, as interested citizens who want to know

the truth? And what can we do to help, or to be prepared when the next "flap" or wave of activity once more fills our skies with something strange?

There are things we can all do. For those fortunate enough to live in the vicinity of Austin, training and participation in PSI activities as a volunteer might be possible. The expense of supporting a research effort such as this suggests a way in which we might contribute. As radio amateurs, however, we have unique qualifications for participating in UFO event-sharing on a national as well as a local level.

We can keep informed through groups which correlate and communicate information, such as the Center for UFO Studies. Dr. J. Allen Hynek, Chairman of CUFOs, was technical advisor for the

production of "Close Encounters." Much of the realism of this film can be attributed to the case information he was able to provide. We can also join or support investigatory groups such as MUFON or GSW, for example. MUFON amateur radio nets meet weekly. On Saturday mornings at 1200 UTC, the 40 meter section meets on 7237 kHz, and the 75 meter section meets at 1300 on 3975.

Every section of the country has investigators trained by some organization to investigate UFO incidents. They are often interviewed by the media. Most of them would be more than happy to speak at a radio club meeting, or to know that local hams are ready to help during a local UFO flap. Many of them need education in the tremendous capabilities amateur radio has for

tracking and reporting sightings and landings. Repeater groups in particular may be interested in learning who to call and what to do if UFOs appear in their area. So the relationship can be one of mutual benefit. Getting qualified investigators to the site of a UFO incident, while it is still in progress if at all possible, is the key to solving the mystery. The government agencies can offer no help, since they have officially declined to investigate UFOs. The police, if they do anything at all, generally report the incident to the Center for UFO Studies via their hotline. Ultimately, news of the incident may filter down from there to the headquarters of one of the investigatory groups such as MUFON. A local investigator is finally informed and hopefully reaches the scene. By then, the UFO and most of the

evidence is long gone. Wouldn't it be much more efficient if hams knew who to contact in their own area to report an encounter? And a call on the 2 meter repeaters in any city ought to furnish plenty of tracking observers or witnesses in a hurry. We can't all have a Project Starlight International in our backyard, but we do have an HT, a mobile, or a low-band rig and know how to communicate. We also have some technical training which helps in describing a UFO and its effects. Working together, we can solve the UFO problem.

Current UFO activity has recently shifted from South America to Australia. The lull in sightings in the U.S. may end at any time. UFOs, the eyes of Texas are upon you! And we'll be keeping ours open, too. ■

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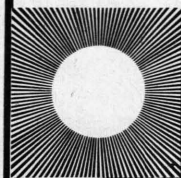
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An X-Band Transceiver

— more 10-GHz fun

The system described in the following paragraphs is made up of surplus microwave components and other equipment used in the home and mobile station here at W1SNN. Only one piece is

described for construction because the main components are now available to amateurs and made for amateur frequencies. The device we will construct is not too difficult to make, but several pieces will re-

quire the use of a lathe.

The method used to control the frequency of the Gunn diode oscillator in the transceiver is not new. It is often used in radio astronomy equipment although in that

endeavor, spectral purity of the rf output is far superior to that achieved here at W1SNN. Many amateurs have used the same idea in days past to control klystron local oscillators for their

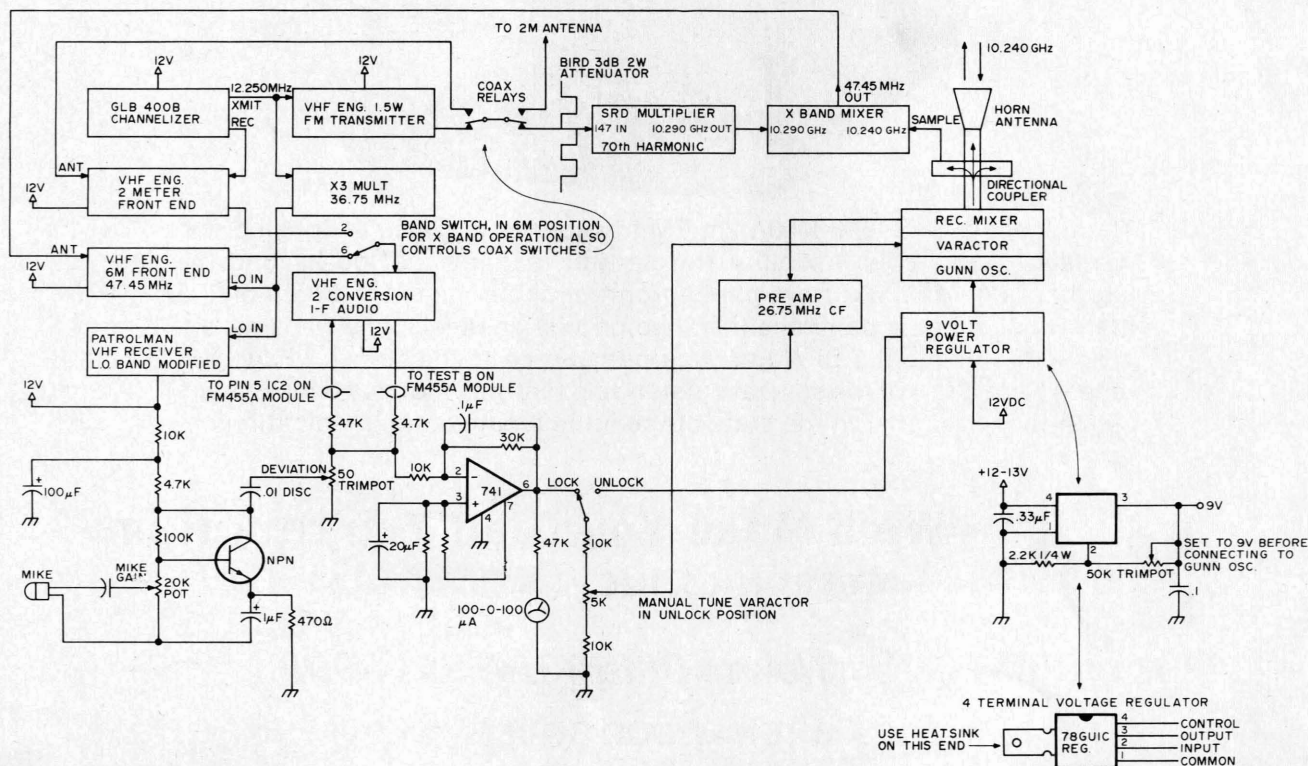


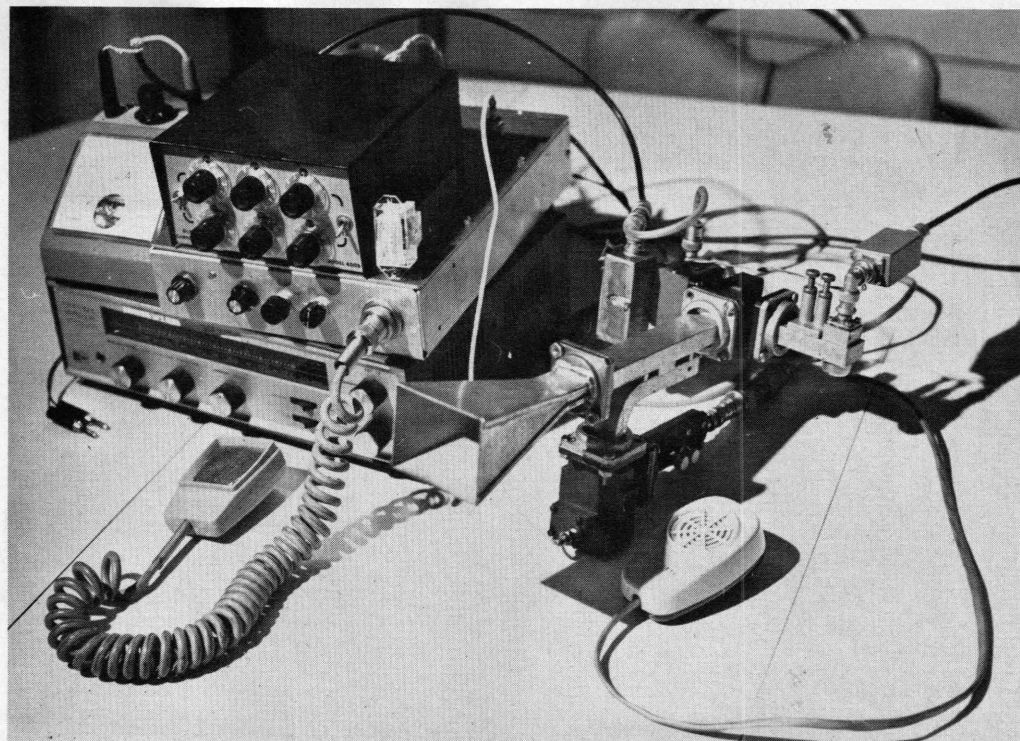
Fig. 1. Block diagram and simplified schematic of 10 GHz synthesizer.

receivers. Let us examine a block diagram of my approach (Fig. 1). I hope you will come up with better ways to do the trick, since we need more activity on 10.250 GHz.

It starts out with a GLB-400B channelizer. The GLB drives a 1.5 Watt two meter transmitter and receiver and an appropriate two meter antenna through two coaxial relays which allow the use of the two meter gear for liason when setting up on X-band during field days. The other output of the second coaxial relay feeds into a comb generator containing a step recovery diode (SRD) or snap varactor which generates the appropriate X-band harmonic.

The comb generator receives its power for harmonic generation by setting the GLB transmit output to 12.250 MHz, which is multiplied up to 147 MHz through the VHF Engineering 1.5 Watt two meter transmitter. The output comb from the SRD or snap varactor produces a usable output at 10.242 GHz. This output is fed into a microwave hybrid mixer and mixed with a sample of the Gunn oscillator's output through a directional coupler to produce an i-f output at 47.45 MHz.

A VHF Engineering six meter front end set at 47.45 MHz serves as the second conversion to the 10.7 MHz i-f input of the two meter receiver i-f amplifier and associated circuits. The i-f input of the two meter receiver can be switched to either the two meter front end or the one on six for liason use, or for synthesizing the X-band transmitter. The six meter LO input comes from the 12.250 MHz output of the GLB through a tripler. The 10.7 MHz i-f is further converted down to 455 kHz where the discriminator output voltage is fed into a servo amplifier. The servo



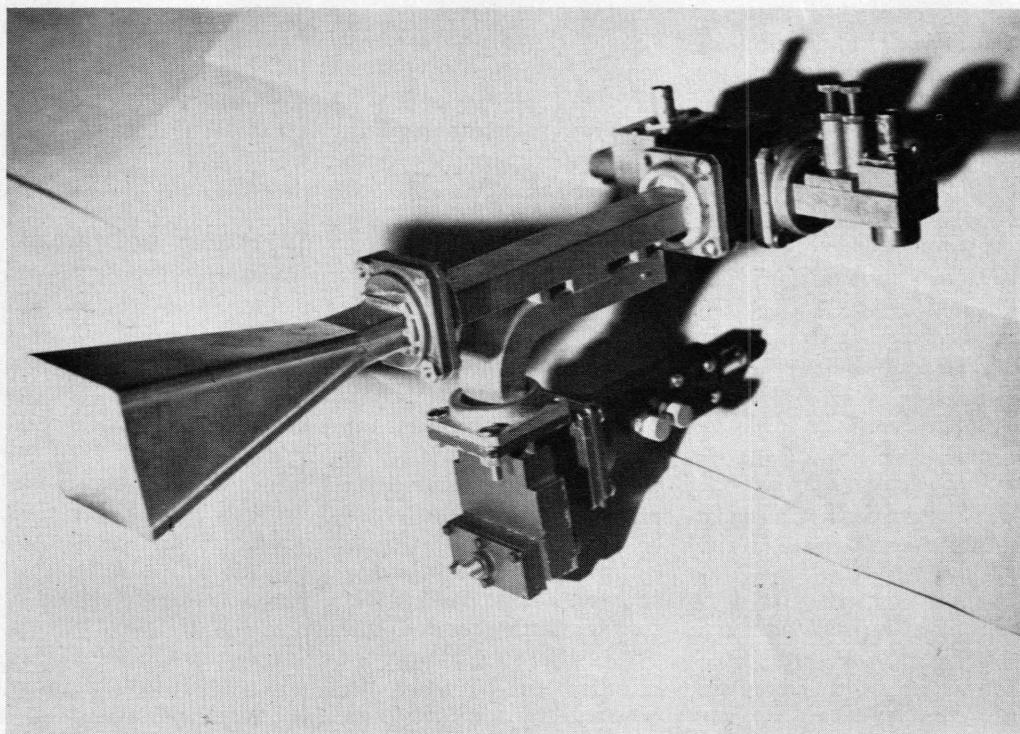
The X-band phase locked rig set up on the kitchen table photo studio. The meter on the left is used for tuning up. The GLB sets on top of the two and six meter receiver. The 1.5 Watt two meter rig is in the same chassis. The box that hangs on the right-hand detector mount is a preamplifier. On the left, a small box contains the servo and modulation electronics.

electronics are complete with an integrator and are used to tune a varactor mounted in the Gunn oscil-

lator cavity. Thus, the system becomes locked to the GLB channelizer.

A second receiver is

made up by modifying an old Radio Shack two-band Patrolman police monitor. This receiver, which is



The rf plumbing stripped of all the electronics. On the right is the receiver detector which feeds into a circulator. On the left is a Gunn oscillator. The large square unit fastened to it is the varactor used to tune the Gunn oscillator.

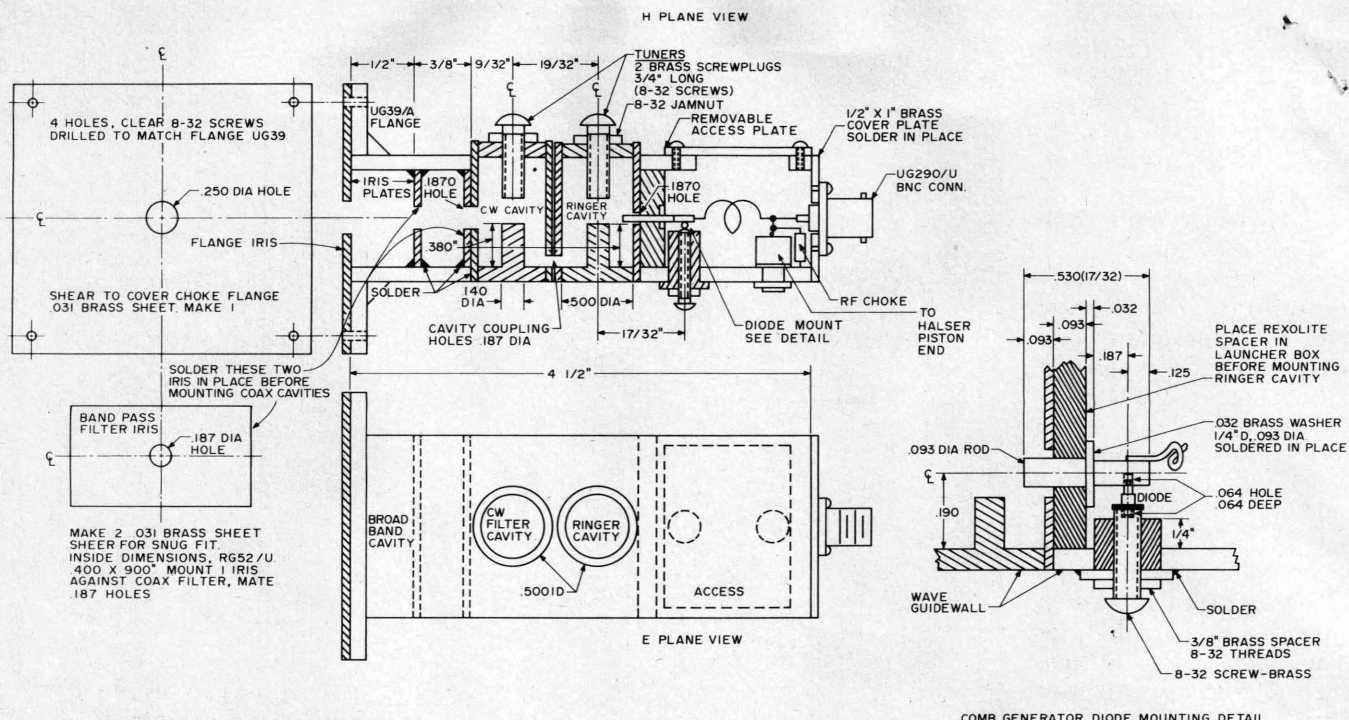


Fig. 2. Harmonic generator construction details.

tunable (or frequency controlled by the GLB synthesizer through the same tripler used above) is used as the receiving i-f system at 26.75 MHz when locked to, or at, frequencies that can be tuned across the X-band range. Modification requires the addition of a preamplifier that would match the output of the receiving mixer to get the noise figure down to a useful level. This receiver does not have any band-narrowing filters in its i-f system (10.7 MHz) and is ideal for the searching that is required for stations that obviously will not be on your frequency. It also has 12-volt capabilities which allow it to work in the car. This receiver could be replaced by another six meter converter and VHF Engineering i-f audio module as is used in the two meter setup. Until this change is made, we will continue to make our contacts with the old police job which so far has given excellent results over the short-path QSOs with WA1NWF. Its bandwidth is wide enough that once the signal is captured, small

drifts can be compensated for by readjustment. It is not an easy task to make contact with another station unless you both are locked on the same frequency, so a tunable receiver, for the time being, is a must. Another amateur, who is using a similar idea, has a police scanner in use as the search receiver. The crystal oscillator has been modified to be tunable and is scanned in the same manner that it would be if it were crystal controlled. This provides a fully-automated scanning system so he can drink his Coors™.

The microwave plumbing used in this set is composed of pieces either gathered from surplus houses or constructed by the author. It is quite a task to make a Gunn-diode oscillator/mixer assembly such as the one used in this system. The frequency control of the oscillator is accomplished by voltage-controlling a microwave varactor located within the oscillator cavity. A ferrite isolator, acquired from a police radar set, and incor-

porated with the tunable crystal mount completes the mixer half of the unit. The construction of these devices is a formidable task unless a well-equipped microwave test facility and a large amount of machine work can be accomplished. Therefore, it would be worth the constructor's time if he incorporated a Microwave Associates Gunnplexer™ which has all three of these items, as well as an excellent horn antenna.

The Comb or Harmonic Generator

Snap varactor or step recovery diodes are names given to these semiconductors by two of the leaders in this field. Undoubtedly there are other names, but, since this article started with these nomenclature, we will keep them. These semiconductors have a property which, regardless of what it is called, performs as follows: An epitaxial diffused varactor is designed to store a charge when it conducts in its forward direction. It conducts for a very short time until this charge is

pushed out by the driving rf signal. Then the conduction ceases very quickly. This is called the "lifetime" and is a way to measure the period that the varactor will store a charge, and the snap-time (or step-time, as it is also called) ceases. These diodes sometimes require an external bias, but ours will be used in the self-bias mode.

The effects described produce a series of pulses, which cause, in our generator, the first cavity to "ring," producing a train of damped waves at a microwave frequency. The output of the "ringer" cavity is then directed into a second cavity, which, by virtue of its high Q, propagates a more CW-like signal to its output termination. In the unit described for construction, these cavities are coaxial for the first two, and waveguide bandpass cavities for the last two. Needless to say, much care will be required in the construction of the unit. See Fig. 2 for details.

The diodes can be obtained from many manufacturers. Two are listed, with references, at the con-

clusion of the article. The two references are certainly worth the reader's inspection in regard to future generator construction.

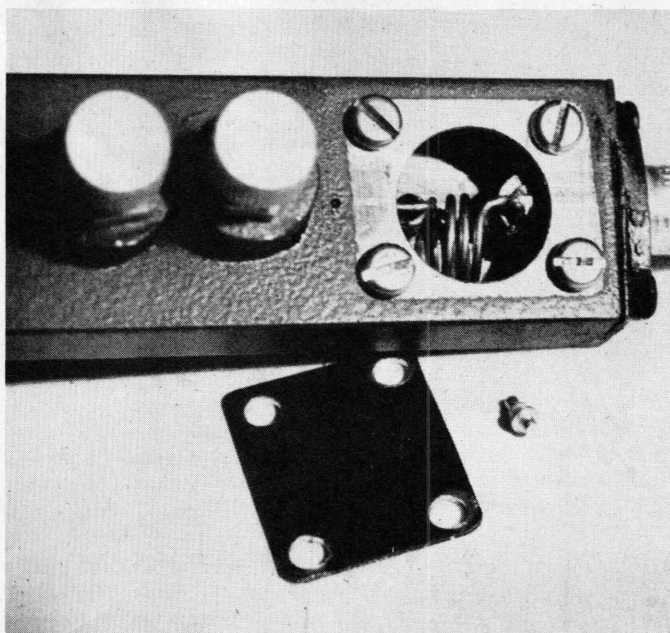
The parts required for this section are easily obtained. They consist of a three-inch section of copper "small X" waveguide (RG-52/U), two 1/2-inch i.d. sections of brass pipe with 1/32-inch wall thickness, and four end-plugs for these pipes that must be machined and drilled to size. Also required are a cover, UG-39B/U, which is soldered to one end of the waveguide, and a small brass box which is constructed from another section of the same waveguide. Construction of this unit should be easily accomplished from the mechanical drawings. The iris coupling consists of holes found near the bottom of each coaxial cavity. The plate used to cover the choke flange, which is the output of the harmonic generator, is the frequency controlling element of the fourth filter, a waveguide bandpass filter.

The tuning screws found on the top of the waveguide must have the large plates loaded to make sure that firm connection is made to the wall of the waveguide. Once adjusted, the jam nuts can be firmly set since no further adjustment will be required, but be sure that the nuts squeeze into the lock washers.

Begin construction of the SRD assembly by marking off the two holes for the circular cavities. These two holes are drilled through the waveguide walls and are spaced double the thickness of the pipe wall, so the material left in the edges of these two holes will be removed. It is best to drill both holes with a smaller drill and then line-ream out the remainder of the material un-

til a tight fit is achieved. The two pieces of 1/2-inch tubing should now touch each other on one side.

Insert each of the cavity tubes into the holes so that the two iris slots near the bottom of each tube face each other. At the center of each of these cavities, the holes shown in the drawings should face the open ends of the waveguide. These are the exit irises of each of the coaxial filters. Next, install the bottom plug which should be the machined piece which makes up the coaxial post for each cavity. All of these pieces should be turned on a lathe so that they provide a very tight fit into the tubing. Make sure that the plug ends of these cavities are parallel to the bottom wall of the waveguide and that the iris holes near the bottom are facing each other. Install the top plugs, which have the tuning studs, into the top of each coaxial cavity. These, too, should be machined for a very tight fit.

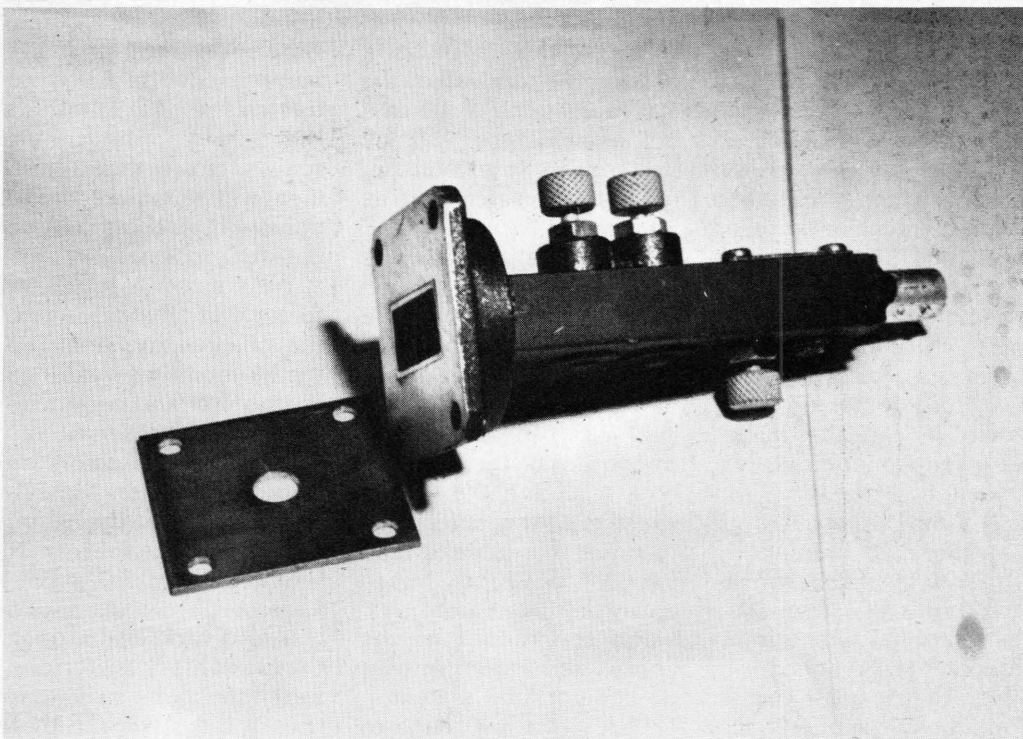


This view shows where the piston capacitor is located. The SRD is in the foreground near the inspection plate.

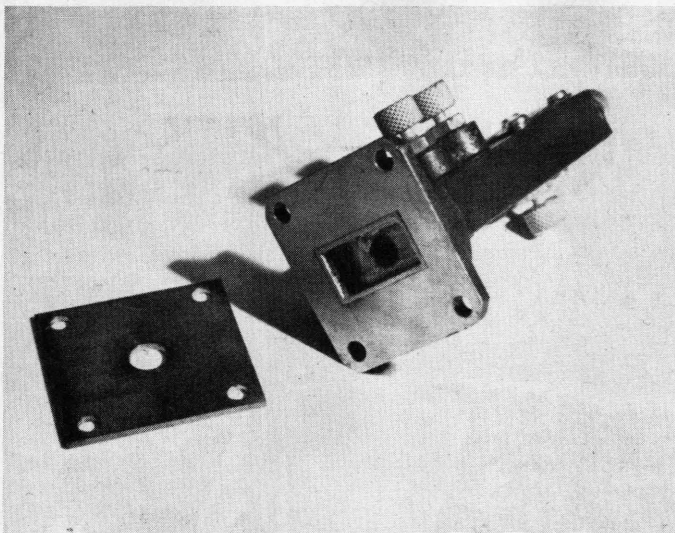
When all of these parts are aligned and you are sure of the correct positioning as shown in the drawing, gently heat the assembly to a temperature that will cause soft solder to run, apply flux to each joint, and run in the solder. Use only a very small

amount of solder; just enough to make a good electrical joint. Let the unit cool before attempting the next step.

On one end of the waveguide, install the inside rectangular iris plates as shown. Then install the output flange on the end of



Side view of the SRD comb generator which provides the harmonics for mixing. The two tuners on the top of the waveguide are inserted into the tops of the coaxial cavities. On the bottom, the SRD mounting stud is in view and just behind it is the piston capacitor tuning plug.



Front view of the comb generator shows the second internal iris which forms half of the bandpass filter. The large flat plate near the flange is the output iris which makes up the second half of the bandpass filter. It is held in place by the mounting screws which fasten the whole unit to the directional coupler.

the waveguide. Again heat the assembly, flux the iris plates and the flange, and apply solder. Care should be taken not to overheat the waveguide such that the cavity assemblies are disturbed. A four-cavity assembly has been completed and is now ready for installation of the SRD and rf launcher.

Examine the drawing of the comb generator diode mounting detail on Fig. 2. This assembly and the driving coil which is tuned to two meters make up the rf launcher. These components must be carefully assembled prior to installing the SRD in place. This diode is a small pill which has a form factor known as "style 31" in the microwave semiconductor trade. To install it, place the end with the large flange near the small circular protrusion into the end of the prepared 8-32 screw, after the screw has been run up through the threaded bushing so that it's just visible. The hole in the end of this screw should accept the diode with a little resistance. Do not push sideways on the diode or you will fracture the seal.

When installed, run the screw into the bushing so that the other end of the diode engages with the launcher line. The launcher should have a matching hole drilled into it in the correct place. DO NOT tighten the diode—just a firm fit will do. All of this activity is accomplished through the inspection hole on the top wall of the waveguide. The diode cannot take much heat, so any soldering of the launcher rod must be done before its installation.

When the assembly is completed, replace the access plate and connect the output of the two meter gear through the attenuator shown in the drawing. This pad will reduce the output of the transmitter to about 0.5 Watts which is all that is required to drive the diode. It can take only slightly more power, so take care!

Put the output of the comb generator into a wavemeter and detector. The detector will have to drive a very sensitive meter if the output is to be seen. There should be at least 20 microamperes output and should be indicated at the

frequency described.

If a smaller indication is shown, it should improve when the tuning screws on the top of each cavity are adjusted. The frequency meter should be left on the prescribed frequency of 10.240 GHz. Adjust the output coaxial cavity, which is the one nearest to the flange, for maximum and then peak-up the ringer. It is possible to be 147 MHz higher or lower since this is a comb generator, so watch the frequency meter. If a spectrum analyzer is available, as was to the author, little problem will be encountered in adjusting for peak output, but if not, it will require judicious observation of the frequency detected by the frequency meter cavity. When it is correctly adjusted, one very large peak right on frequency can be measured. When it is slightly off peak, a number of peaks will be observed when the frequency meter is tuned through the desired output. With a little practice, full output will be assured. When adjustments are completed, be sure to tighten the two jamb nuts that are on the tuning screws. No further adjustment will be required if it is done right as these nuts do not work loose.

We are now ready to try to lock up the system to the channelizer, hook up the harmonic generator to the system, and determine that all of the servo connections and indicators are in order. Turn on the system and look at the center scale tuning meter. It should be right on the line. To prove it is locked, and will also lock again, open the lock/unlock switch and watch the meter swing one way or the other and then snap back to center when the switch is closed. With the lock switch closed on the search receiver, you should be able to hear

yourself when you speak into the mike. Full duplex can be used on this frequency.

The circuitry shown for the power supply, servo electronics, and modulator are simple and should give little trouble. The output from the VHF Engineering i-f strip discriminator should be used as recommended from the manufacturer for driving a microammeter. The meter, however, is replaced with the circuitry shown, and then the meter will be connected to a new set of connections.

Modulation of the varactor requires a little care. Very little modulation is needed as full FM is used. If you are working a station that uses a wideband receiver, then a greater swing will be required. However, if you are working one that has a system like the one described, then the deviation must be adjusted to fit his receiver using the deviation control.

Tests throughout the fall have gone on with this rig. It is portable, by virtue of its several boxes, and gets hauled up on Prospect Hill in Waltham on Sunday afternoons. Prearranged contacts have been made over 30-mile paths with little difficulty.

Plans for another rig of the same type to be used for expeditions are in the works. It would be interesting to see what can be done at greater ranges, which I know are done commercially. It's quite a thrill to hear full quieting on top of a hill with a lot of old junk plumbed together. I hope you will enjoy the same results on your own expeditions. ■

References

Hewlett-Packard Application Notes #920 and #928
Step Recovery Diode, HP 5082-0830
Microwave Associates Snap Varactor, MA 43004

Building Long Yagis for UHF

— some pitfalls to avoid

Beware of the "something for nothing" offer! With this thought in mind, the long yagi antenna is being placed under suspicion. Would you believe a 48-element collinear array (24 driven elements and 24 reflectors), whose gain is approximately 16 dB, could be replaced with a single 13-element long yagi? In addition, the long yagi has: 1. Less wind

resistance; 2. More directivity per pound of aluminum; 3. Greater ease of fabrication; and 4. Excellent mechanical stability. Certainly there is a temptation to believe that the long yagi antenna offers something for nothing when compared with other antenna arrays. What must be sacrificed to obtain the aforementioned advantages? Followers of the

"Collinear Clan," read on!

The cautious amateur constructing a beam antenna intended for application in the VHF and UHF range would exhibit a tendency to cut the elements longer than necessary. This stems from the old adage, "It is easier to cut 'em long than to add a piece on." This is the philosophy that has helped give the long yagi a questionable reputation.

In an attempt to reinstate the long yagi in its rightful position, antenna patterns of a 13-element yagi were made. Fig. 1 shows polar patterns of the long yagi in horizontal polarization. Fig. 2 shows patterns with the yagi vertically polarized. Fig. 3 shows the reference dipole in the horizontal plane. Curve A is plotted in the same relative gain scale as

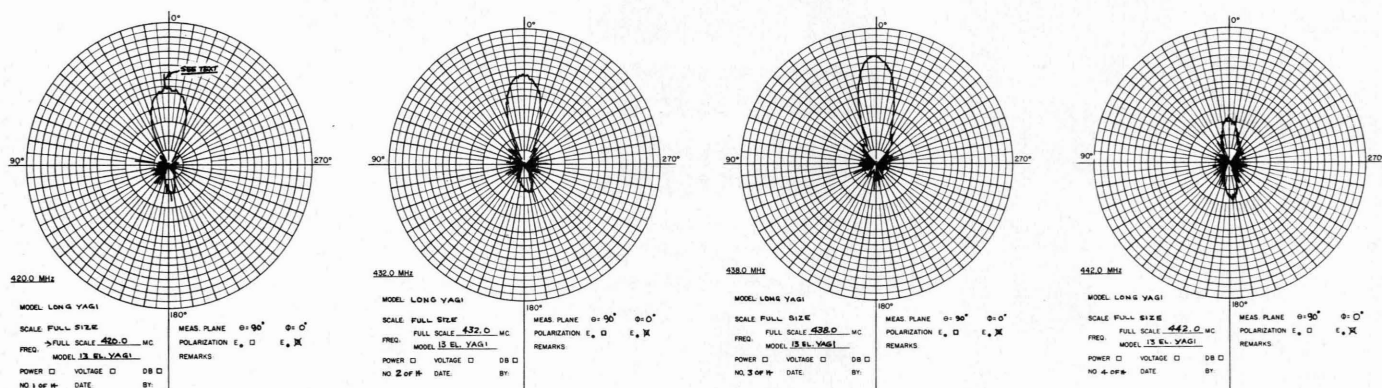


Fig. 1. Horizontal polarization radiation patterns for the 13-element yagi at 420, 432, 438, and 442 MHz.

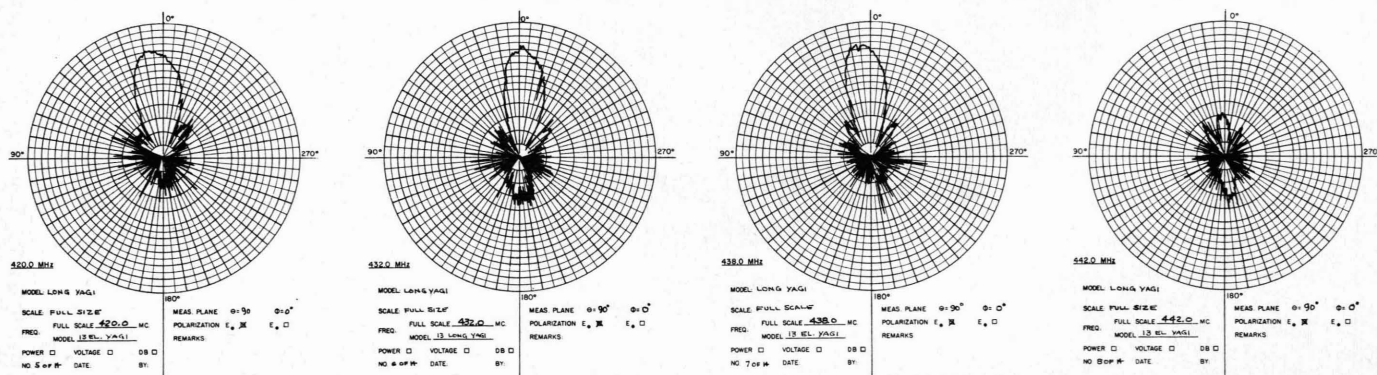


Fig. 2. Vertical polarization radiation patterns for the 13-element yagi at 420, 432, 438, and 442 MHz.

Fig. 5 shows graphically the nonsymmetrical band-pass characteristics of the long yagi-type of antenna. The extremely rapid cutoff on the high side of the intended operating frequency has caused many an amateur to cast aside this type of antenna. This curve shows readily that it is unwise to cut the antenna to operate below the intended operating frequency.

Four long yagi antennas (Fig. 6) operating on 432 Hz, using 13 elements each, will yield a gain of 22 dB (assuming 16 dB per each 13-element yagi). To obtain this same gain with a collinear array, approximately 176 elements would be required (88 driven

Fig. 3. Reference dipole radiation pattern in the horizontal plane.

elements and 88 reflectors). At this point in comparison, does the singular feature of unnecessary bandwidth seem worthwhile?

If reasonable care is exercised during construction (Fig. 7) and *all* dimensions are adhered to within one sixteenth of an inch, then you can boast, "Have vagi. will radiate." ■

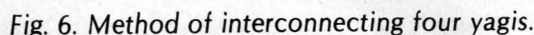
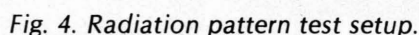


Fig. 7. Construction details for the 13-element yagi.

A Complete X-Band Transmitter

—easy to build

Microwave experimenters in the X-band often require a signal source that has a means of adjusting the output amplitude and frequency. Frequency calibration is also desirable. The above characteristics can be found in standard signal sources, most of which are war surplus and have disappeared from the surplus dealers' shelves or from laboratory stock rooms which had friendly hams operating them. To be sure, these items are still available from micro-

wave manufacturers, but they're at prices well beyond the amateur's means.

The construction of a signal generator which can have laboratory features is not too difficult. The requirements for laboratory standards are much more stringent than most amateurs need, so some of these features can be omitted.

At W1SNN, I constructed a signal source which serves two purposes. It serves as a signal source and as a low-power

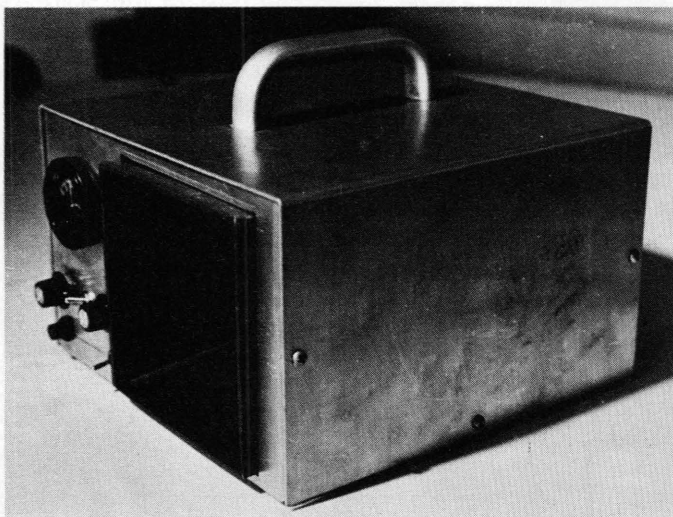
transmitter. It has an internal power supply, a modulator, a frequency control, an output attenuator, and a means of measuring the frequency on which the oscillator is operating.

A klystron tube is used because I had it in the shack. A Gunn diode oscillator could be substituted without too much trouble, although the modulator would have to be considerably modified. Klystron tubes are still found on the surplus market. These tubes, for the most part, are the type which were used as receiver local oscillators. The 2K25, also known as the 723a/b, sold for as little as 50 cents and is still available, as is the BMI series. The 2K25 makes an excellent signal source, but it must be operated upon to get it in the 10.5 GHz amateur band. The BMI series uses a heated bellows to tune to the frequency desired; the bellows heat up, expand, and change the cavity resonance. It therefore is limited in its desirability because of the attendant tuning/stability problems. I

was lucky enough to find a Varian X-13-B, which tunes through the band of interest by adjusting the reflector voltage and fine tuning the cavity with a micrometer barrel which is attached in a manner that allows the cavity to be distorted via pressure exerted from the barrel.

The description of the operation of this and many other fine signal sources can be found in the references given at the conclusion of this article, so I will address myself to the business of construction of the signal generator. The choice of the tube must be yours, but this article will be centered around the 2K25 klystron because it is the most easily acquired tube. The price of this tube is still within the means of the amateur, even for a new one. It is, of course, no longer sold for fifty cents, but what is these days?

The signal generator/transmitter is composed of five sections: power supply, modulator, oscillator, attenuator, and rf indicator. In the set to be described here, a transmitting horn is incor-



An X-band signal generator/transmitter, klystron oscillator power supply, and all components required.

porated into the same chassis. This can be omitted if you wish. The output of the generator can then be terminated in an output flange.

The power supply is very straightforward, consisting of a full-wave rectifier and filter designed to be free of ripple. It consists of two sections. The main supply supplies +300 volts at 75 mA. A capacitor input filter choke capacitor combination is used in this supply. The second supply is for the reflector of the klystron, which also requires good filtering, but, since the drain on this element is in the microampere region, a simple RC filter is used. Heavy filtering is used on each section to improve on the short-term stability of the klystron oscillator, which is voltage-tuned. Vacuum-tube regulators could be incorporated to improve upon this last feature but would not greatly reduce the drift in the oscillator tube, which is also temperature-sensitive due to the cavity construction. This idea was not attempted and is not recommended unless the added expense is your bag.

The modulator consists of a single transistor amplifier, which will give more than adequate modulation. The reflector of the klystron is modulated by the amplifier, and, since the reflector current is very small, equally small modulating currents are required. For the most part, the modulation will be FM, depending upon how hard the tube is pushed. A panel control to determine the modulation level has been incorporated.

Perhaps the part of the unit most difficult to construct will be the klystron tube mount, the level attenuator, and the rf output detector which is used to determine the output level. This group of components is constructed on one

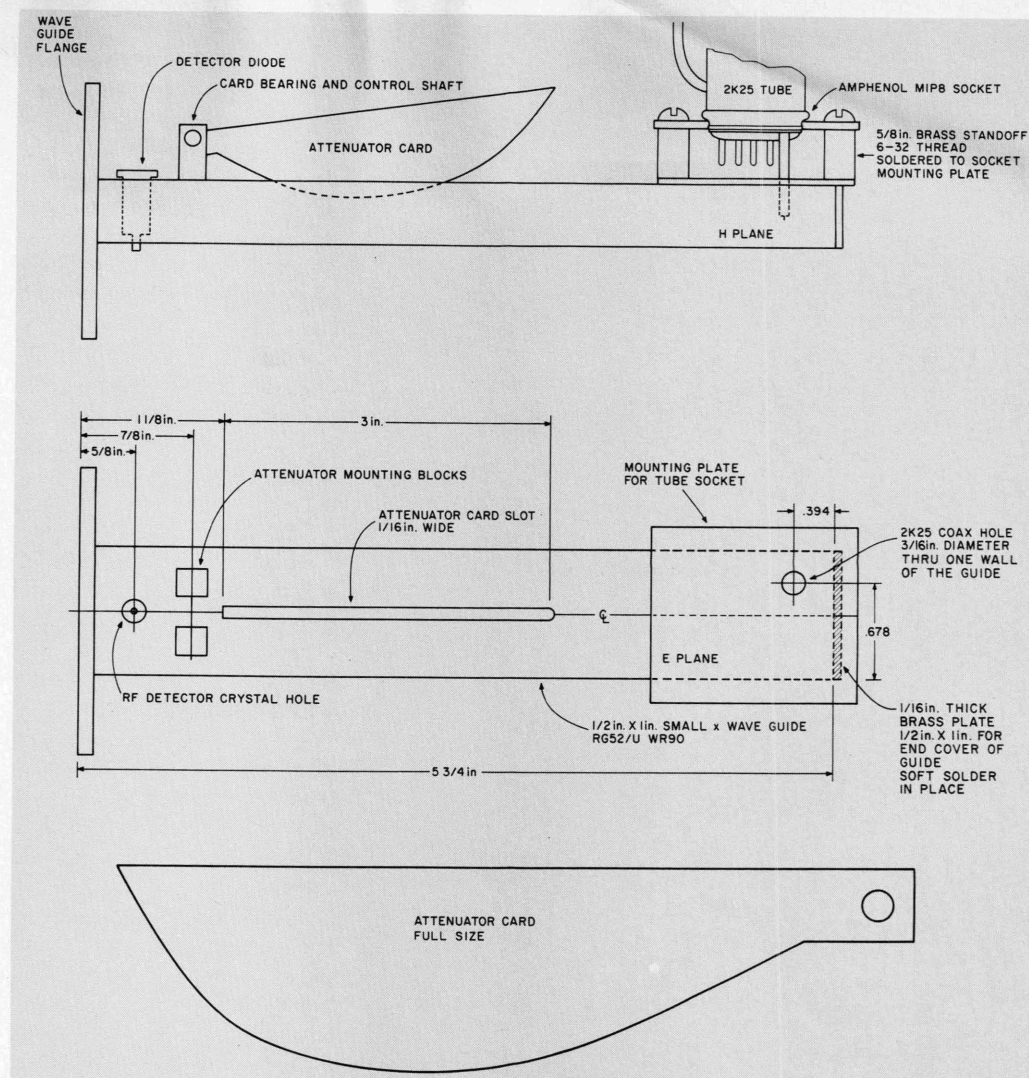


Fig. 1. Waveguide tube mount for the 2K25 klystron attenuator and rf level detector.

single piece of waveguide. No machine shop work is required. All of the work was done at W1SNN with hand tools in the cellar workshop.

The first part of the rf assembly will be the construction of the attenuator. It is a "flap" attenuator, consisting of a section of resistance card that dips into the waveguide through the center of the E-plane. These attenuators were common on the surplus market, and, if one is available to you, by all means use it. The attenuator at this station uses a right-angle gear drive so that the control dial can be conveniently located on the main panel. It may be brought out of the side of the chassis, if

you prefer.

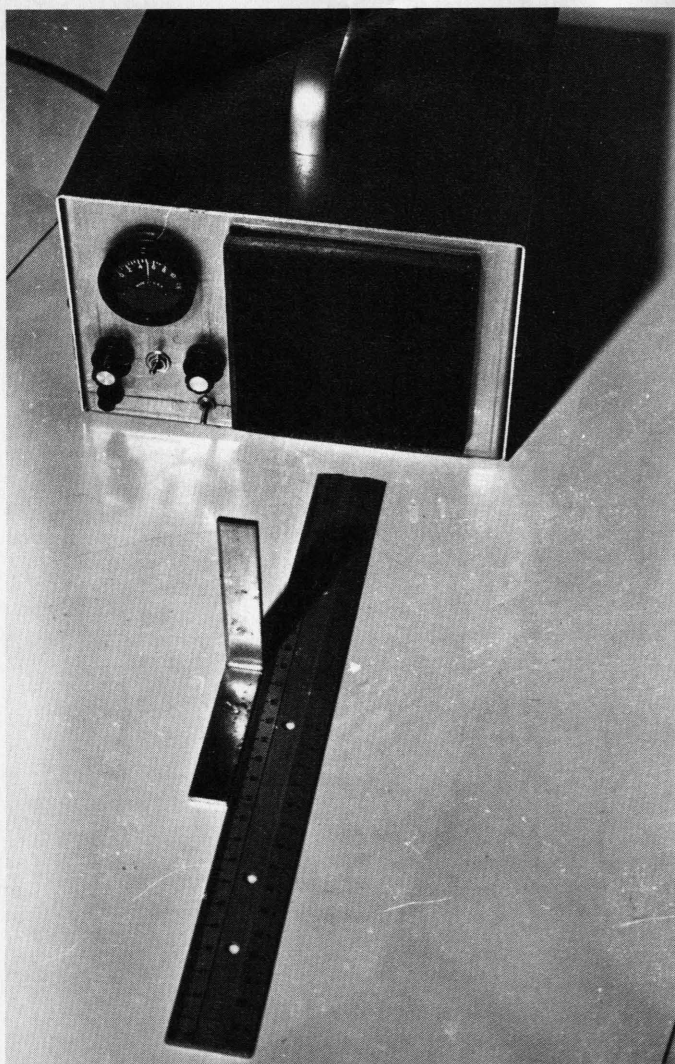
The rf indicator consists of a crystal detector mounted near the output end of the waveguide. A potentiometer and a microammeter connected as a voltmeter to the crystal complete the detector. This equipment may be omitted if you prefer. It is easily constructed and is very useful in making comparison and frequency measurements. Its exclusion, therefore, should be carefully considered on the basis of price only.

The antenna incorporated in this set is a pyramidal horn described in an earlier issue of 73,⁴ so the construction details will not be given in this article. The horn antenna can be substituted for with

whatever gain standard horn is available, saving a great deal of the construction work.

As is shown in the drawings and photographs, the chassis is packaging for the unit so that it can easily be transported. No blower is incorporated because the klystron does not require additional cooling. However, if a different device is used in substitution for a 2K25 tube, be sure to determine if it requires external cooling.

Let's get on with the construction of the waveguide assembly. Acquire a six-inch section of waveguide. This is a little longer than required, but, by the time the ends are squared off, it will be just right. First, with a machinist's square, mark



Frequency measurement—not really the hard way.

off each end so that the scribe lines will show intersecting marks on all sides of each end as square corners. The distance between these should be $5\frac{3}{4}$ inches. Cut the ends carefully with a hacksaw on the lines. File each cut smooth. Remove all burrs, and clean these cuts so that they may be soldered. Steel wool or sandpaper should be used for this purpose.

This first pair of cuts to the guide determines the overall length and the position of the waveguide coupling hole for the klystron. Be sure that the length is that which is specified in the drawings, or you may have difficulty acquiring full power output from the klystron.

Next, lay out the line

through the E-plane center of the guide which will be cut as a slit $\frac{1}{16}$ of an inch wide. This slit allows the piece of resistance card to enter the waveguide. To cut the slit carefully, center punch ten places in line on the centerline so that the punch marks are very close to one another but not touching. Then drill ten holes with a number 43 drill. By careful manipulation of the drill, these holes will be in line and can be opened to start the slit. A hacksaw blade can then be inserted in the cut and the slit further lengthened to 3 inches. This sounds like a lot of work, but once the holes are in place, it goes very fast. When the slit has been completed, insert a flat jeweler's file, and

smooth file the cut, removing burrs. Be sure the burrs that extend into the waveguide are removed.

The next hole to drill is the crystal mount for the rf output detector. Actually, it consists of two holes. The drill should go through both walls in the E-plane. Use a $\frac{1}{8}$ " drill. When this operation is completed, the hole on the same side as the slit should be opened up to $\frac{5}{16}$ ", which will accept the large end of a 1N23 crystal.

The remaining $\frac{1}{8}$ " hole should be fitted with a small piece of $\frac{3}{32}$ " brass pipe to fit the tip of the crystal. The pipe can be soldered to the waveguide with soft solder after it has been determined that the alignment of the holes is true. Use a crystal to prove this point. The $\frac{3}{32}$ " brass pipe can be obtained at a model shop.

To mount the crystal for the level detector, first place a $\frac{1}{4}$ " lug on the crystal body and slide it up to position on the large end. It will come up snug against the flange. Bend the lug so that it can be soldered. Now fit a $\frac{1}{4}$ " fiber shoulder washer on the crystal body, snug against the lug. Be sure that the shoulder faces the tip end of the diode. Lay this assembly to one side until all soldering is completed.

The attenuator mounting consists of two $\frac{1}{4}$ " square blocks $\frac{3}{4}$ " high, soldered at the position located in the drawing so that they straddle the attenuator slot. A $\frac{1}{8}$ " hole is line-drilled through the sides of each block, which then can be fitted with a small axle for the attenuator card. The card is then epoxied to the shaft. A $\frac{1}{8}$ " to $\frac{1}{4}$ " coupling is attached to the axle so that a control shaft which fits a conventional dial can be added.

When this job is com-

plete, solder a $\frac{1}{2}$ " x 1" plate over the end of the waveguide which is closest to the tube socket for the klystron. On the opposite end, solder in place a UG-39/U waveguide flange.

To complete the klystron mounting, it will be necessary to modify an octal tube socket to fit the klystron tube. Inspect the tube, and note that it has a small piece of coaxial cable made of metal tubing protruding from the pin end of the tube. One end of the coaxial cable has a small length of Teflon™ covering a short section of the inner conductor of the coax. This is the coupling antenna used to inject the rf output of the klystron into the waveguide. The coupling depends upon the location of the entry hole in the wall of the waveguide and the depth that the antenna extends into the waveguide. The octal tube socket should have a mounting assembly attached to it. Usually it is held in place with a spring washer or is molded in place. The specified socket is still listed in catalogs, since tubes have not completely disappeared from the scene. Choose a socket, and remove the contact from pin 4. Open the pin hole to accept and freely pass the coaxial output of the klystron tube. Some sockets will have a hole large enough and will not require modification. The socket used at W1SNN is an Amphenol type MIP8.

The tube socket is mounted to a small plate approximately $1\frac{1}{2}$ " square. Solder two $\frac{5}{8}$ " threaded bushings in a position that will allow the antenna probe of the klystron tube to clear the hole in the waveguide wall intended for it. A similar hole will have to be added to the mounting plate. Once the location is determined, which will depend on the socket type, the

*The attenuator card can be constructed from a piece of resistance card made for that purpose. It is beyond the scope of most amateurs to try to manufacture this component. Its cost is reasonable and it goes without saying that it should be purchased.

To cut the resistance card, simply scribe the outline as shown in the drawing, being careful not to disturb the surface of the card. Then, with a sharp knife, cut through the scribe marks. When the cutting process is complete, file the edges smooth and to the contour shown in the drawing. Departure from the shape shown will cause the attenuation rate to bunch up near the lower levels. The card is quite thin, but not as fragile as it appears. The card should clear the slot as it enters so that wear on the resistance material will be nonexistent.

The schematic diagram illustrates a vacuum tube radio receiver circuit. The power transformer has a 115V 60Hz primary and secondaries for 680V CT 70mA, 6.3V AC 2.5A, and 5V unused. The power section uses a 2N3404 transistor for rectification and a 723AB/2K25 Klystron for RF output. The AF section includes a 2N3404 transistor and a 723AB/2K25 Klystron. The circuit is powered by a 115V 60Hz transformer and includes various components like capacitors, resistors, and a 50K potentiometer for RF level indication.

guide work has been completed, install the components, making sure that the tube socket wiring is completed before installing it. Then decide what kind of cabinet or container you will use, if you decide not to use the one shown in the drawings. Make clamp mountings for the waveguide, as shown, and construct the power supply and attenuator control. The meter and level setting for it can be installed, and you will be ready for testing after a slight modification is made to the klystron bellows.

Simply rotate the square brass tuning control, which is fastened to the tuning strut, full clockwise until the strut opens to approximately 3/16". This action

Install the klystron carefully into the socket, making sure that the antenna probe slides into the hole in the waveguide so that it just exposes the end of the coax.

preliminary adjustment. The unit pictured and shown in the drawing is complete with a horn antenna, since it is used as a transmitter. It makes a fine portable setup for testing waveguide and other components, since the horn is easily removed at the flange. In any case, connect a horn radiator to the output of the waveguide.

41

justing the reflector control. Three positions on the reflector control will cause the indicator to rise sharply as you "tune" through the klystron's modes of operation. Each of these modes will electronically control a small deviation in frequency of about 100 megahertz, so get used to how these modes occur, and become acquainted with the level potentiometer setting for midrange for each mode. Now exercise the attenuator control, and note that the output level can be raised or lowered to completely control the rf output.

Once you have familiarized yourself with the operation of each control and the way the klystron acts, it will be time to set the frequency. If you have a waveguide frequency meter which is calibrated, choose one of the reflector modes—preferably the middle one—and adjust the control for the "middle" of the mode. Adjust the attenuator so that attenuation is minimum and connect the wave meter to the output flange. Adjust the wave meter for a reaction on the wave meter indicator. Track the wave meter with the reflector control to determine where 10.200 GHz falls. If you find that you must move to the third mode, readjust the two nuts on the tuning strut, as well as the brass tuning block, starting with the block first. Once the frequency is located, carefully make a mark on the reflector dial for this frequency. Then find the limits of the band, and mark it again.

If you do not have an X-band wave meter at your disposal, you will have to resort to the method that the early microwave people used. It is as accurate as you use it and will certainly locate the band limits for you.

To set up for this opera-

tion, you will need a small plastic ruler which is marked off in metric measure. A small piece of aluminum one-inch wide and bent in the form of an L will serve as the reflector. Set the signal generator/transmitter up with a horn on a wooden or nonconducting surface with a clear area in front of the horn antenna. Measure a distance of 15 cm from the lip of the horn. At this point and through the axis of the horn, lay the meter ruler. Tape it in place so that it is firm and will not move. The ruler is the calibration source and should be treated as the standard for these measurements.

Now that the ruler is fastened to the work surface, place the L-shaped bracket next to the cm side of the ruler. Tape a pin to the bracket so that it points to the calibrations on the ruler. Fasten the pin so that it won't move. Now proceed to fire up the generator as before, and locate the middle mode by adjusting the reflector potentiometer so that the mode is peaked on the rf indicator meter. Move the L-shaped bracket parallel with the ruler so that it bears on the ruler. Move the bracket toward the horn and away from it a few times, and notice that the rf indicator fluctuates rapidly as you move the bracket. It will also fluctuate as you move your hands or any other object in front of the horn.

Now move the L-shaped bracket, parallel to the meter ruler, toward the horn from about the middle of the ruler. Remember that your hands will influence the readings, so try to hold them flat to the work surface as you move the bracket through a minimum and a maximum reading on the rf indicator. Be sure to note the reading on the ruler; the distance between these two points

is one half the wavelength of the transmitted frequency. Make a series of these measurements to determine that they repeat, recording the distances. You already know how to convert wavelength to frequency. All you have to know is that the physical free space length is .2997 to .2855 cm for the frequency range of 10.000 to 10.500 GHz. As before, if the klystron does not operate in these frequency ranges, it is a simple matter to change the coarse adjustment, as previously described.

If you want to know more about the way to measure the wavelength on which you are operating, I suggest you find an old *Radio Amateur's Handbook*. The issues prior to 1955 had a good amount of data on lecher wires.

Since a lecher line is really an open wire transmission line, attention must be given to the correction in length, which is usually foreshortened by capacitance and inductance between the wires. Although this factor is small, it is there, and it affects the accuracy of the measurements. In the free space case, which is the method I use, there will be no problem with the foreshortening effect. Simply find the two points where the reaction takes place, measure the distance between them, and you have the measurement as a half wavelength. As you can see, the measurement must be very carefully made. All components in use must remain stationary except the L-shaped reflector. If this part could be moved with the precision of a metric micrometer, the accuracy would be improved far beyond your needs. It is important that you establish the band edges, since other services surround you; the police operate at 10.525

GHz $\pm$ 25 MHz, and you don't want to interfere with the police radar frequency, do you?

The modulator in the signal generator/transmitter is a single transistor amplifier that is used to amplify a microphone, audio oscillator, or whatever you decide to modulate with, except video. For the most part, modulation is FM. Deviation is dependent upon the setting of the gain control.

You may want to investigate the possibilities of using a Gunn diode oscillator in place of the klystron. There are several available, complete with horns, if you have the money, or you could use the unit described in this magazine as the "Mobile Smokey Detector."³ It's as easy to build and test as is most microwave gear found in 73.

If you decide to construct this unit and use it on a field day for another band multiplier, look for me on 10.250 GHz. A parabolic reflector six feet in diameter makes the output of this unit formidable. The QRM up there isn't too bad from other amateurs, but it is quite often experienced from radar. The band is a lot of fun to work on—hope to see you there. ■

References

1. *The Microwave Engineer's Handbook*, 1963, Horizon House, Brookline MA.
2. *Technique of Microwave Measurement*, Radiation Laboratory Series, Volume 11, Chapter 2, "The 723 Family of Reflex Klystrons."
3. "Mobile Smokey Detector," Stirling Olberg W1SNN, 73 *Magazine*, Holiday, 1976.
4. "A Brass Horn For X-Band," Stirling Olberg W1SNN, 73 *Magazine*, March, 1978.
5. The attenuator card material is available from these manufacturers: Filmohm Corp., 48 W. 25th St., New York NY; Emerson and Cummings, Inc., 869 Washington St., Canton MA; Film Resistors, Inc. (HyTronics), 242 Ridgedale Ave., Morristown NJ.

New Life For Double Sideband?

—awake, ye pioneers, and get cracking

In the 1930s, when a ham spoke of "radiotelephone," he meant only one mode — good, old-fashioned, full-carrier AM. When hams (or FCC exams) discussed alternate types of modulation, the choice was between grid and plate modulation; FM and SSB were laboratory esoterica for the most part confined to the future.

Since then, most amateurs have learned the folly of full-carrier AM. Single sideband predominates on the low bands, while FM is the workhorse of VHF. Few voice operators have had any experience with another mode that combines the best of FM and SSB with the simplicity of "Ancient Modulation."

What's AM, Anyway?

The popular view of modulation is fraught with misconceptions.¹ Many hams still believe that "amplitude modulation" is accomplished by varying the strength of a "carrier wave" in step with the modulating signal. This is erroneous; by definition, the carrier of an AM signal is not changed by the modulating process and carries no intelligence. In fact, it doesn't carry anything; it just sits there. Audio and rf are mixed in the modulated stage of an AM transmitter, producing sum and difference frequencies which are called sidebands.

The sum and difference products are redundant, and

the full modulating signal can be found in either one of the two sidebands. In SSB transmitters, the carrier and one sideband are removed. This is perfectly sufficient, given the proper reception techniques.

An AM detector, whether a simple diode or an SSB product detector, is nothing but a mixer. The carrier frequency is mixed with the sidebands (one or both), producing difference products which duplicate the original modulating waveform. If the carrier is not transmitted, a beat frequency oscillator (bfo) fills its role within the receiver, mixing with the SSB signal. See Fig. 1.

The nominal signal bandwidth of an SSB signal is that of the modulation. Since the usual voice spectrum is from 300 to 3000 Hz, an SSB signal is 2.7 kHz wide. An AM signal is twice as wide as the maximum modulating fre-

quency, so voice AM uses 6 kHz. This is one of the most important reasons why SSB has supplanted AM on the crowded HF bands.

AM has another big problem — heterodynes. Since AM is just mixing, two nearby carriers can mix and produce audible difference products. CBers are today plagued by heterodynes, as were hams twenty years ago.

Perhaps AM's biggest disadvantage is the energy wasted in the carrier, which must always be twice as powerful as both sidebands together. But it also has one big advantage, which few hams (or anyone else, for that matter) make use of.

Redundancy — FM's Secret Weapon

"Frequency modulation" is often believed to be a process of shifting the carrier's frequency in step with the modulation. This isn't any truer than the carrier-strength theory of AM discussed above. In FM, the carrier sits on a single frequency, but it is surrounded by multiple sidebands, whose phase relationships give the appearance of varying frequency. At one point in the modulating cycle, the upper sidebands will tend to cancel, while the lower sidebands tend to reinforce one another; at the other half of the cycle, the phases switch position. The amplitude of the carrier does vary; a constant amount of power is divided between the carrier and the sidebands.

FM is capable of better noise rejection than AM and has the "capture effect" — a stronger signal will completely obliterate (capture) a weaker signal on the same frequency. While FM's noise resistance is a product of the phase relationships, the rest of its benefits are primarily due to redundancy. The wider the FM signal, the more sidebands; the signal-to-noise ratio rises with bandwidth. Unfortunately, the wider receiver bandwidth needed for

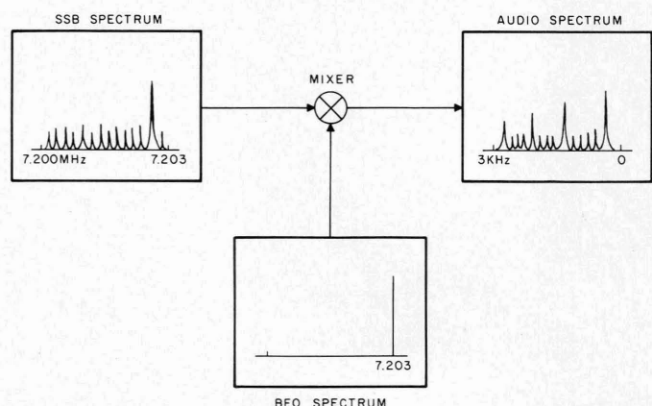
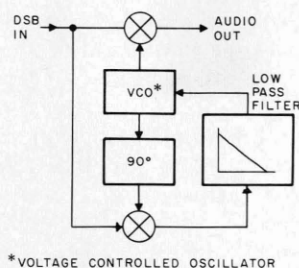


Fig. 1. SSB spectrum is mixed with a carrier in the detector; the output is the difference. In an AM system, the bfo is not needed, since the transmitted carrier is mixed with the sideband(s).



*VOLTAGE CONTROLLED OSCILLATOR

Fig. 2. PLL detector for DSB.

The only special tools required are a vice, hacksaw, and a file to cut the Zephyr down to size. The Zephyr is disassembled from its base and the chrome parts are re-

The pendulum is cut down, and the existing hardware is moved around (see Fig. 2) to convert to the electronic keyer paddle con-

It took about four hours to modify the Vibroplex Zephyr and would probably take about the same time to modify other kinds of bugs. Visitors to the shack will be impressed with the good

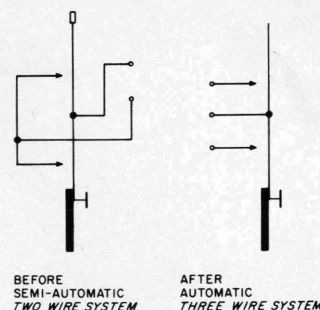


Fig. 3.

looks and low cost of this professional-looking shack accessory. ■

The circuit is a basic NE555 timer circuit with the addition of a reed switch. I was unable to find a relay with a low current consumption, and was able to use the reed switch in its place. The total cost of the entire project was less than \$4.00 using all new parts.

The push-button switches are used to reset the timer and pull in the reed switch initially, to provide power for start-up. I used two switches, but a single DPST normally open push-button switch could be used. The only disadvantage of using one switch is that there is no way to shut the unit off before the timing period is up. The switch connected to pin two of the NE555 will shut down operation if pushed before the timing period is up. An SPST switch could be placed in the battery line to provide this feature if so desired.

The reed switch is a General Electric number GE-X7 Experimenter with

With the CW filter on, the circuit draws 22 mA. This should provide a normal battery life. Whatever is connected to point "A" should not draw more than 250 mA. ■

Don't Let Your Battery Die

—*extend its life*
with this simple timer

Sam Miller WB8TXG
4893 Timbercrest Dr.
Canfield OH 44406

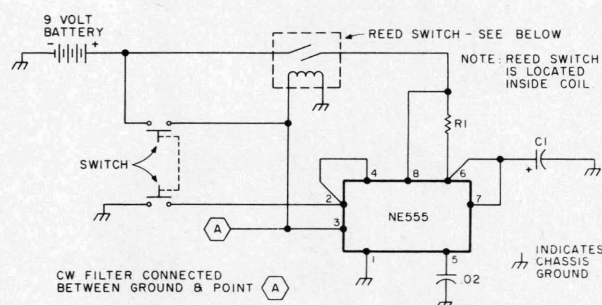


Fig. 1. R1 — 5 meg. C1 — 50 uF electrolytic. Switch — either 1 n.o. DPST push-button or 2 n.o. SPST push-buttons. Reed switch — #GE-X7 Experimenter line by General Electric with 7,000 to 10,000 turns #36 enamel covered magnet wire around switch form.

wideband FM admits more noise, so FM requires a stronger signal than SSB to overcome receiver noise.

An FM detector always takes advantage of redundancy. But regular AM has some redundancy, too — two complete sidebands. How can we put this to good use? A diode detector won't suffice, since it mixes all signals indiscriminately. A product detector only works with SSB. To get the most from AM, a synchronous detector is necessary. And a good synchronous detector doesn't require the presence of a carrier between the two sidebands. It enables us to use double sideband, a mode almost forgotten by history.

A Quick History of DSB

Double sideband without carrier (DSB) was developed in the mid-1950s. General Electric developed military communications gear using DSB, but the Collins Radio Company, with its SSB gear, beat GE for the pace-setting military contracts. Some have attributed this to better lobbying on Collins' part, although their equipment certainly can't be slighted for quality. Since then, DSB has been mostly ignored. A few ham DSB transmitters came out around 1960, but they were intended for hams who couldn't afford SSB; the other guy wasn't supposed to notice the "wrong" sideband!

Unlike SSB, DSB cannot be received on an AM/CW receiver. A synchronous detector is necessary. The only way to receive DSB on an SSB receiver is to filter out one sideband. A synchronous detector isn't a simple device. It revolves around a phase locked loop (PLL). How many hams knew about PLLs in 1956? The new PLL chips aren't designed for DSB, but they can be used in DSB receivers. An optimal DSB detector, sometimes called biaural or bisynchronous, requires 20 tubes or so, but it becomes quite manageable with ICs. A single chip could

be built for it, but it hasn't been yet.

Synchronous Detection

The basic difficulty with DSB is that the bfo must be exactly in phase with the carrier used to generate the sidebands. Absolutely perfect stability is necessary, unless phase locking is used. Happily, phase locking isn't tough to achieve. So, by injecting a phase locked bfo into a product detector, DSB can be received. The basic DSB detector in Fig. 2 just keeps the bfo in phase. The result of this is a system that is at least as efficient as SSB and doesn't require any expensive crystal filters.

A true biaural detector makes full use of redundancy. Its operation is basically simple. If a signal isn't simultaneously present in both sidebands, it's spurious, so reject it. This may sound like a nifty trick to pull out of a hat, but it was described in 1956 by John P. Costas W2CRR and, later, in a 1966 73 article — with lots of tubes!

Fig. 3 is a block diagram of the biaural detector. It has two parallel signal paths — the I channel and the Q channel. The incoming DSB signal is fed into two product detectors (balanced mixers). The bfo is fed directly into the I channel. It is shifted 90° (by a resistor and capacitor) before being fed into the Q channel. An adequate 90° phase shift is easy to produce over a narrow range of frequencies (less than an octave at a time), and it requires no adjustments if the detector is used across an entire ham band in a direct-conversion (synchrodyne) receiver.

The I (in-phase) channel will detect everything present in the input, including both the desired DSB signal and any unwanted signals and noise. But the Q (quadrature) channel won't. If you combine a DSB signal with its carrier shifted 90° (in quadrature), you'll get phase modulation, not the original AM.

That's how many FM transmitters work. But, since phase modulation (practically the same as FM) can't be heard on an AM detector, the Q channel will not detect the desired DSB signal. It will hear everything but the desired signal. So a detector just has to shift the Q channel audio 90° (with no great precision; you don't need the expensive shift networks used in phasing-type SSB transmitters) and subtract it from the I channel. The desired signal will remain — free from QRM!

The Q detector has another function. Its output, combined with I in a phase detector (double-balanced mixer), produces a dc control voltage that locks the bfo onto the desired signal. As the oscillator or signal drifts, the signal appears in Q with the right polarity to correct the bfo.

So a DSB transmission-reception system combines the weak-signal performance of SSB (even allowing for the wider bandwidth) with the frequency-correcting ability of FM with automatic frequency control and adds a unique ability to suppress QRM.

Frequency Overlap

If two SSB signals have overlapping passbands, a receiver will hear them both. One will be distorted, but it will still cause interference to the other. But, with properly detected DSB, the phase locked signal will be heard, and the overlapping signal will be attenuated. How ef-

fective this would be on crowded ham bands remains to be seen, since hardly anyone has tested DSB lately, but 10 dB of QRM rejection seems a conservative guess. Signals may overlap without interfering — wouldn't that help in pileups?

On VHF, DSB would prove particularly useful. There's no shortage of band space, and DSB is just as good as SSB for weak-signal DX. Since SSB transmitters for VHF are quite complex, simple "plate modulated" DSB transmitters — AM transmitters with balanced finals — could enable many more hams to enjoy the DX potential of 6, 2, and above.

DSB Transmitters

Short out the crystal filter in a typical SSB transmitter, and it will put out DSB. That's not the easy way to do it, though, unless you have a spare SSB exciter or two. Fig. 4 is a simplified schematic of a DSB final stage. It's quite like an AM final, if you can recall that many years ago, but the carrier is balanced out.

Fig. 5 is a sneaky circuit that uses that most modern of components, the MOSFET. It's just like a plate-modulated AM stage, but it doesn't put out a carrier. A MOSFET has an extremely high-input impedance, like a class A tube amplifier, and it also has a low-output impedance, like a tube. But amplifier tubes are essentially rectifier tubes with one or more control elements. A triode tube is just a diode with a

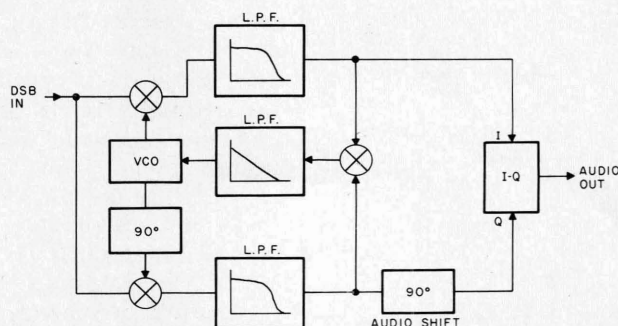


Fig. 3. Biaural detector providing rejection of undesired signals.

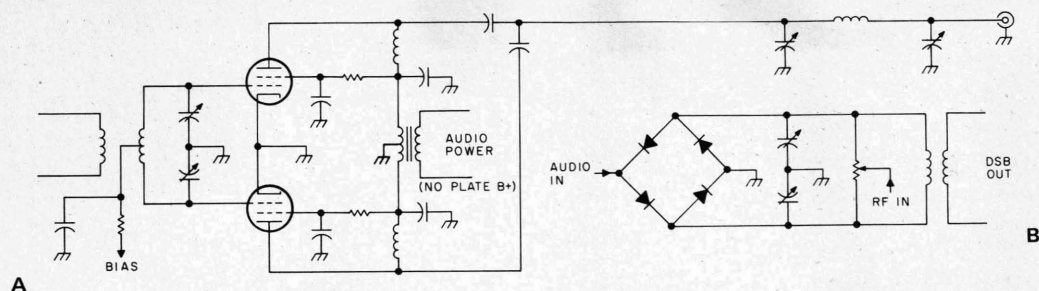


Fig. 4. (a) Balanced modulator using tubes; (b) Ring-type balanced modulator for use in low-power stage of a DSB or SSB transmitter.

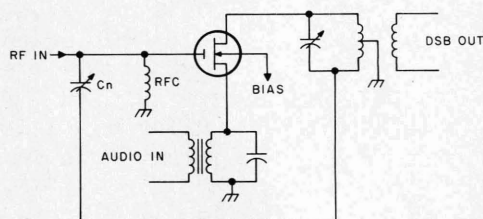


Fig. 5. Single-device balanced modulator. FET must be bidirectional. VMOS power FETs and many others are not; check spec sheets.

grid. FETs, on the other hand, are not diodes. The source and drain (cathode and anode) can be reversed, and current will still flow. Control is by the voltage between the gate and the source.

Notice that this modulated amplifier does not have any source of dc. All power for the stage comes from the audio amplifier. In the absence of audio, there's no output. What could be more balanced than that? By properly neutralizing out the input capacitance of the FET, up to 90 dB of carrier suppression is attainable.

A tube final amplifier turns dc into carrier and generates sidebands from the audio power. If the negative voltage peak of the audio exceeds the dc supply voltage, the tube stops conducting over that portion of the audio cycle and distortion (splatter) results. This can't happen with a bidirectional FET, which stays linear below 0 volts. So, even if you apply dc in order to produce a carrier, there will never be a splatter. Any amount of carrier can be used, from zero on up to the transistor's own limit. But the intelligence-bearing sidebands are not affected. Even the AM boys

should like that kind of splatterproof rig!

Synchrodyne Reception

The cheapest ham receivers today are the direct conversion, synchrodyne or homodyne, units that mix an oscillator with the input signal, detecting either SSB or CW. No i-f is needed. The receiver oscillator doubles as the CW transmitter oscillator in one such rig, the popular Heath HW-7. But, on SSB, the bandpass of the receiver is twice as wide as desired. The 3 kHz below and above the oscillator is heard, but that means you hear undesired signals along with the one you want.

Since DSB makes use of both sides of the carrier frequency, a synchrodyne has the perfect bandpass. A synchronous detector automatically tracks frequency and phase drift on a DSB signal, and it doesn't require fixed selectivity filters. So a synchronous detector can be operated over a wide frequency range as a direct-conversion receiver. No i-f strip or conversion is necessary, which makes for the best possible dynamic range and sensitivity. The cost of the detector is compensated for by the saving in the i-f

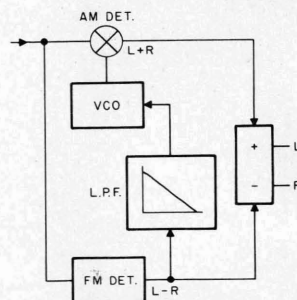


Fig. 6. Simplified AM stereo receiver. Everything but the output network is part of a standard PLL circuit.

stages and filters. Selectivity comes from the low-pass audio filtering in the detector.

AM Stereo

The big money in the radio broadcast game has been shifting to the FM band in recent years. The noisy, low fidelity monophonic sound of AM broadcasting just can't compete with FM where both are available. But AM broadcasters are planning a comeback, thanks to some of the techniques outlined above.

Synchronous detection permits AM receivers to reject heterodynes and much other noise (not, unfortunately, pulses such as ignition noise). Higher fidelity AM is thereby made possible. But, even better, it makes AM stereo practical.

Two rather similar systems have been proposed for AM stereo. Neither is a simple left-on-the-upper, right-on-the-lower sideband system. Instead, the two channels (L+R) that constitute a regular monophonic AM signal are generated using double-sideband techniques with sup-

pressed carrier. The carrier that was used to generate the DSB is then phase modulated with the difference (L-R) audio. An AM stereo receiver then has an AM detector drawing L+R from the sidebands, and an FM detector drawing L-R from the transmitted carrier and its associated sidebands. See Fig. 6.

Two of the systems presently being considered by the FCC differ primarily in that one, proposed by Leonard R. Kahn (an SSB pioneer), has a 90° phase difference in one modulator, whereas the other, proposed by RCA, does not. But that's another story...

An AM stereo receiver doesn't require the sophisticated phase locking circuitry of a DSB receiver, since it has a carrier to phase lock to. Recall that PM doesn't actually change the frequency or phase of the carrier — modulation always consists of adding sidebands. ICs like the 561 PLL detector can be adapted to such a system, while existing mono AM receivers won't even notice the difference. Unfortunately, DSB without carrier can't be received with such inexpensive chips, which are really exalted carrier AM detectors, not synchronized to the sidebands. The chip makers haven't built them, yet.

Too few amateurs understand the virtues of two complete sidebands. While conventional full-carrier AM is terribly inefficient, balancing out the carrier and using a synchronous receiver results in a communications system with weak-signal efficiency at least the equal of SSB, with substantial QRM rejection, automatic drift compensation, and reasonably economical hardware. Amateurs should give more thought to using DSB, especially for long-haul VHF work. ■

Reference

1. Goldstein, F.R., "AM Is Not Dead — It Never Existed at All," 73, May, 1976, p. 110.

The Op Amp Beam Heading Indicator

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Robert C. Ghormley KØBV
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I saw this linear ohmmeter circuit in the Nov. 11, 1976, issue of *Electronics*, although its theory and operation were treated lightly.¹ My particular application was readout of antenna rotor position using only two wires

connected to the rotor potentiometer (one end plus center tap; the other end of the pot was not connected). The pot resistance between the two wires increased linearly from zero resistance to 1000 Ohms, corresponding to 360° of rotation. Any application where a linear readout of a variable resistance is desired may be filled by this circuit.

If you just want to build

this circuit, are going to go buy all parts, and don't care how it works, use the following: R1 equal to the maximum value of the unknown resistance you're measuring; R2 = 2.2k; V_Z (zener diode voltage) = 3.3 V/400 milliwatts; a 0-1 mA meter; R3 = 2.7k; R4 = 1000 Ohms; a TO-5 can 741 op amp; and a +12 V power supply. Skip toward the end of this article for calibration procedure and cautions. For those who will use junk-box parts, the following will let you roll your own.

A garden-variety 741-type op amp is connected as in Fig. 1. This circuit uses a single power supply, with the op amp's negative supply terminal grounded. The non-inverting (+) input is biased up to a regulated level (any convenient zener up to 5 or 6 volts will function okay). This serves to bias this input and to get the op amp operating "above ground." The inverting (-) input is returned

to ground through a resistor equal in value to the maximum unknown resistance to be measured. The unknown resistance, RX, is connected as a feedback resistor.

The output of the amp feeds a current meter (anything from a 50-microamp to a 1-milliamp movement will work just fine) and a series current-limiting resistor. Note that the negative meter terminal is returned to the reference voltage (the non-inverting input), not to ground. The series variable resistor allows full-scale calibration.

Pick a meter with four major divisions — 0, 25, 50, 75, 100, or something like that. Relabel these S, W, N, E, S, depending on where you wish the antenna to be indexed. To do a neat-looking job, disassemble the meter and erase the existing legend with an ordinary pencil eraser. This takes a little time, but it works (eraser crumbs inside the meter movement are to be avoided). Relabel the scale with dry transfer letters and reassemble.

Any supply voltage from +10 volts to +30 volts, or even slightly higher, will work. Too much will pop the op amp, but they're cheap. The zener should be biased on by R2 so that it draws 3 or 4 milliamps or more (anything above that is just wasted power). C1 and C2 are bypass capacitors (.001 to 0.1 μ F disc ceramics will work fine).

In operation, this circuit functions as an inverting amplifier. The gain is $-(R_X/R_1)$, which varies from zero to minus one as R_X varies from zero to R_1 . The "signal" voltage which is being amplified is the difference between the left-hand side of R1 (ground) and the noninverting input (the zener potential). If this zener voltage were +5.1 volts, for example, the output of the op amp would vary from 5.1 volts, for $R_X = 0$, to 10.2 volts, for $R_X = R_1$. The output of the amp is the reference voltage (zener voltage)

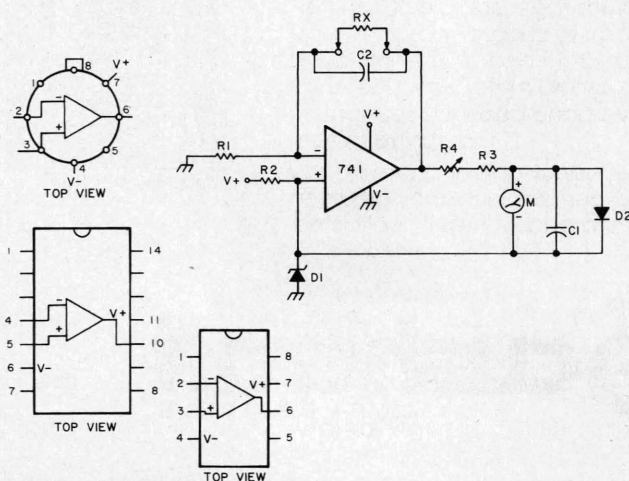


Fig. 1. Linear ohmmeter.

plus the amplified "signal" voltage. If RX is allowed to approach infinity (open terminals), the gain approaches the open-loop op amp gain (a large gain indeed), and the output of the op amp will approach the positive supply voltage, pinning the meter. If this happens often, add the meter protector germanium diode D2 (1N270 or equivalent); it will still pin, but it won't smoke. The series resistors, R3 and R4, are selected to produce a full-scale meter deflection when the op amp output is maximum.

To calibrate, let $R_X = 0$ (a short). The meter will show no current (the op amp output voltage will equal the noninverting reference voltage). Now clip in a resistor for R_X which is equal to whatever you want the full-scale meter reading to correspond to (1000 Ohms in the case of my TV-type rotor). Adjust R4 to obtain a full-scale meter deflection. Now, just to prove to your-

self that it works, cut R_X in half, and note the half-scale reading. Letting R_X equal zero (shorted) will produce no meter deflection.

Selection of the zener diode should be made with one eye on the supply voltage. At full-scale meter reading, the output of the op amp will be approximately twice the zener voltage. If a 9 V zener is selected, the maximum op amp output will be 18 V; hence, a power supply voltage of at least 2 or 3 volts above this should be used. A 12 V or 24 V c-t transformer feeding a bridge or full-wave rectifier with a couple hundred microfarads on its output will yield about 18 volts and work just fine. Reference voltage (zener voltage) could then vary from 3 volts up to 6 or 7 volts with no problem. No power supply regulation will be required, unless you're powering something else from the supply and drawing a lot of current.

If the unknown resistance

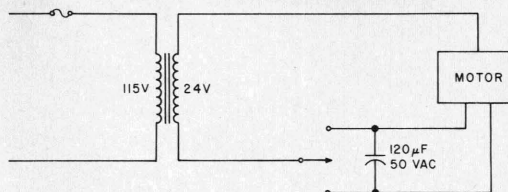


Fig. 2. Rotor motor drive scheme.

is connected to earth ground (or chassis ground) at either end, the circuit will still work if the power supply is completely floated. One end of the pot in my rotor was grounded, so I just floated the negative power supply lead above chassis ground and connected it only to the op amp circuit.

For those who are interested in fabricating the complete control head for a Cornell-Dubilier-type rotor, I offer the following. I have three different rotor types which all use the same driving scheme, shown in Fig. 2. C1 is a 120 μ F/50 V ac rotor replacement capacitor. If this capacitor goes bad, the rotor will turn slowly and erra-

tically, or will refuse to start. Try replacing it before you dismantle the antennas and bring the rotor down! Make sure the switch is a momentary center-off type. Alternatively, you could use a normal toggle switch there with a momentary push-button single-pole switch between it and the transformer. This scheme will not tell you that you have reached the rotor stop, except by the meter indication. If a momentary switch is not used, it would be easy to burn up a rotor motor by leaving power applied. ■

Reference

1. "Direct Reading Ohmmeter," V. Ramprakash, *Electronics*, November, 1976, p. 115.



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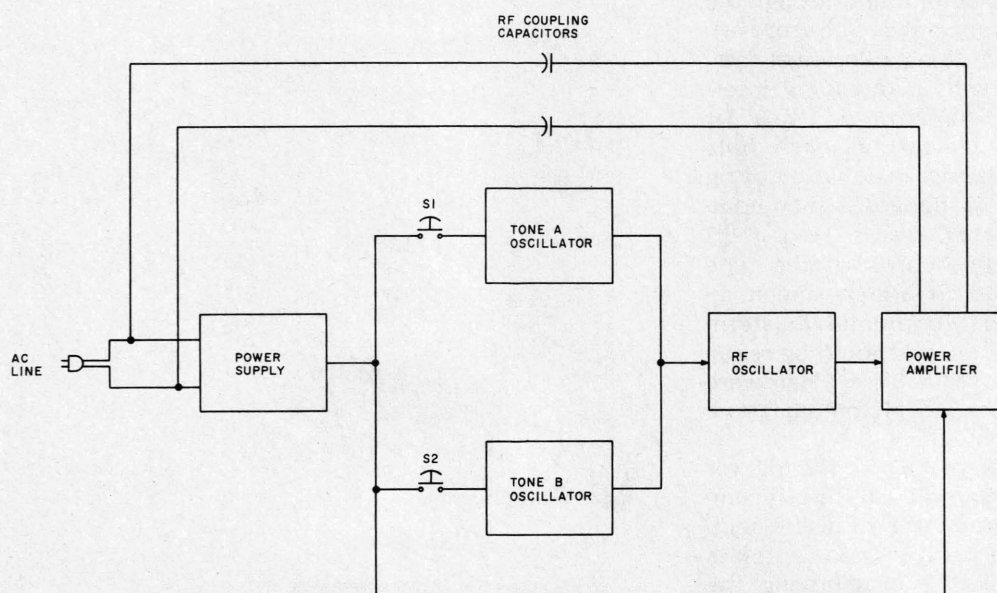


Fig. 1. Transmitter block diagram.

Like many hams, I'm interested in radio as more than just a medium for yakking. Radio frequency signals can be employed for a variety of things far beyond ham-type uses. This article shows how "wired wireless" (WW) signaling can be used for remote control functions.

The system shown here is used presently for a rather mundane function. It controls lighting from a remote location external to a house. However, it could be used to switch almost anything electrical or electronic within or outside a house. The only limitation is that there must be a common power line between the transmitting and receiving circuits. Such functions as controlling an antenna switching relay right at the antenna, turning on and off ham shack equipment from anywhere in the house, or even turning on a coffee-pot from the bedroom come to mind immediately. I'm sure you can think of many more applications. Additionally, the circuits included in this system can be used individually for wireless intercoms, burglar alarms, or remote monitors.

Getting Interested

As I hinted above, WW has fascinated me for quite some time. My first exposure was via the so-called wireless intercoms. Naturally, the "wireless" part is not strictly true (just like the British term wireless used for radio). There are interconnecting wires between the intercoms, but no extra wiring. They operate by transmitting rf signals through the 110-volt power lines. Each unit is like a low-power transmitter/receiver combination. Rather than transmitting electromagnetic waves through space via antennas, the signals are conducted through the power line. The useful part of sending signals this way is that there is no need to worry about extra wiring or antennas — the signal path is free.

Aside from wireless inter-

goms, there are numerous other devices using the same or similar techniques. Many burglar alarm installations, where additional wires are awkward or impossible, use WW signaling. Heathkit sells a home alarm system using plug-in sensors and a central unit designed along these lines. Lafayette Radio and others also sell remote-control devices for home appliances that signal via the ac line.

Those of you who live near modern schools may have noticed a funny whistle in your TV or radio at the same time that the school bells ring. This, too, is a form of WW system in which a master control sends coded signals through the school's power wiring to control the bells. Some of them also set the school's clocks exactly with another set of signals.

Large power companies have made their high-voltage long-distance transmission wires serve double duty. They send supervisory and voice communication information on rf carriers along with the low-frequency power. Increasing usage, too, is being made of WW signals for load shedding to turn off non-critical electrical appliances during peak power demand times.

Digging Deeper

Although information about these various systems is usually scanty, I devour any piece of it I can find. Wireless intercoms introduced themselves to me through a friend's request to repair a nonworking pair. The operating frequency (about 200 kHz) was listed on the label, but very little additional information was. Naturally, I also traced out the schematic. As luck would have it, the heart of the matter was an assortment of inscrutable rectangular coils in small metal cans bearing the inscription "Made in Japan."

Similarly, I once ran across some wireless remote controls sold by Lafayette Radio.

Tracing out the schematic diagram showed that the basic circuit was extremely simple. The transmitter portion was an ordinary one-transistor oscillator that ran somewhere between 100 and 200 kHz. Two oriental coils were used, one as a tuned circuit in the oscillator and the other apparently as an rf transformer used to match a transistor output amplifier to the ac line.

The receiver also used two transistors and two coils. The first was an amplifier connected to the power line via a link-coupled tuned transformer. Output from this amplifier was coupled through a second tuned inductor to the second transistor. As I remember, the second transistor acted as a detector and power switch to turn on a relay when rf was present at the power frequency.

Heathkit uses a similar system and circuitry for their home alarm system. Each sensor unit (fire, smoke, door switch, freezer warning, etc.) has a special transmitter module consisting of an astable multivibrator operating just below 100 kHz. The multivibrator is fed with half-wave unfiltered 60 Hz dc power, causing the output to be rf pulse amplitude-modulated at 60 Hz. It sends rf through the power line through a power resistor connected to one side of the multi which draws extra current for half of each rf cycle.

In the Heathkit receiver, there are a couple of tuned rf amplifiers (again with special inductors) which feed an AM detector. The detector picks 60 Hz modulation off the incoming carrier. A second detector then senses this 60 Hz audio and turns on an SCR and relay.

Other commercial wireless burglar alarm systems use similar techniques. However, they must be more selective and reliable, so they require better selectivity. Some I've

seen use ceramic resonators in both sending and receiving units. These resonators behave like the more expensive quartz crystals used in radio communications. They enhance the transmitters' frequency stability and give the receivers good selectivity to discriminate against noise and stray signals.

Still another class of equipment that uses power lines for signal transmission is home entertainment equipment. About ten years ago, a few record changer/tuner combinations were sold with remote speaker capability. There was a small modulated oscillator connected between the record changer and the power line. The remote speakers each had a small receiver which intercepted the signal from the ac wiring and fed it to an internal audio

amplifier and speaker. Although I never learned the circuitry of these gems, I did learn that they used an FM signal between 400 and 500 kHz. In fact, there was also one high-quality system that transmitted stereo via two different carriers between 400 and 500 kHz.

General Characteristics

Now let me summarize a few things I learned while investigating the previously mentioned systems. First, they use the power lines for their transmission medium. Second, they usually use frequencies between 50 kHz and 500 kHz.

Transmitter power levels are usually only a few milliwatts in home use, although the power companies may use several Watts for their long-distance systems. I measured receiver sensitivities of about

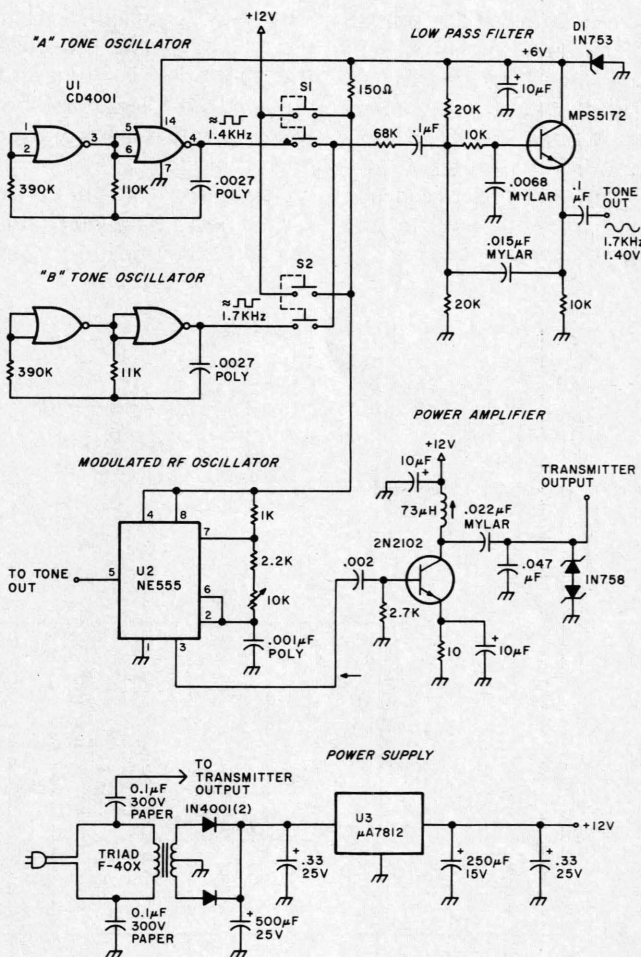


Fig. 2. WW transmitter schematic. Notes: (1) Polarized capacitors are electrolytic or tantalum. (2) Poly means polystyrene. (3) Unspecified capacitors are disc ceramic.

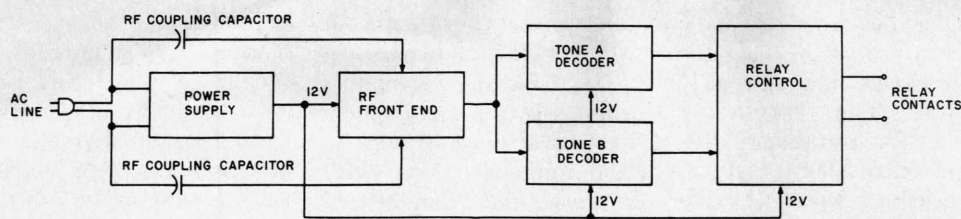


Fig. 3. WW receiver overview.

1 mV for the most sensitive receivers. Simple systems use AM, but FM systems are more reliable and noise-free.

The carrier frequencies in the above range are chosen for several reasons. They are high enough to handle as radio frequencies as far as the transmitters and receivers are concerned, but not so high as to require special circuitry. Very importantly, they are easy to transmit short distances via power line conduction, but are not radiated well because of their long wavelength. Power company pole-mounted transformers attenuate them so that the signals rarely are sent from one circuit to an adjacent one. The receivers usually aren't subject to interference problems, since services in this frequency range are not powerful and they are few in number.

The legality of most WW systems is somewhat hazy. Section 15.4(d) of the FCC rules seems to come the closest in defining such a system as a "restricted radiation device." Then Section 15.7 gives the limits for allowable field strength. Rf level must be below 15 microvolts/meter at a distance of 157,000/F (kHz) feet from the apparatus. They likewise require that no interference occur with a licensed service. And, finally, the equipment must bear a certificate of compliance with Part 15 regulations based on measurements by a qualified technician. Of course, the wording they use is somewhat obscured in "legalese," but what I've said is a pretty good interpretation.

Regardless of the cited regulations, I have yet to find any such "certificate of com-

pliance" on any commercially sold equipment I've seen. Whether this is intentional on the FCC's part or just what the story is I'm not sure. At any rate, if you do any experimentation with wired wireless, be sure not to use too much power or do anything else to interfere with a licensed service. Taking these precautions preserves the intent of the FCC regulations and should keep you from attracting their attention.

The Project Begins

In spite of my continuing interest in and occasional use of commercial WW equipment, I'd never really built my own. About five years ago, I did try a few circuits in an AM intercom, but the results were terrible. I couldn't couple enough rf into the power line to overcome the noise generated by fluorescent lights, dimmers, and electric motors. I was never sure whether the problem was insufficient signal or poor receiver design. Perhaps it was a little of both.

Then about two years ago, a friend asked me to help with a problem he had. Basically, he wanted to

control a lamp post at the far end of his driveway, but there were complications. First, there was no power available at the post when the light was off, since the switches were inside the house. Second, he didn't want to run any extra wiring, since it would have to be buried and it is about 200 feet from the house to the lamp post (surgeons don't live on postage-stamp lots).

Because of the no-extra-wire limitation, radio control was an obvious solution, but the distance was rather long for an unlicensed transmitter system. Of course, the lack of available power was a real handicap! I began thinking of a wired wireless transmitter as a possible solution. But this still had two drawbacks. There still was no power to run equipment at the light post, be it wireless or otherwise. More importantly, when the power was off, the transmission path for the rf signal was broken.

Both these limitations, however, were easy to get around. Since the desired control point for the lamp was right at the lamp, a battery-powered WW transmitter could be installed. Nickel-cadmium batteries could be used, connected so that, when power was on to run the lamp, they would be trickle charged. To get the signal around the switch in the house (when off), it was bypassed for rf. Since it was an SCR switch, a series-connected 0.1 uF capacitor and 100-Ohm resistor were connected across its terminals.

The capacitor has low enough reactance at rf to pass the control signal. At 60 Hz, though, the reactance is high enough to limit current flow to about 5 mA. The 100-Ohm resistor protects the SCR from being destroyed by current surges caused by the capacitor.

I built the system, and it worked very well. I seriously doubt though, that many 73 Magazine readers want or need such a control. So what

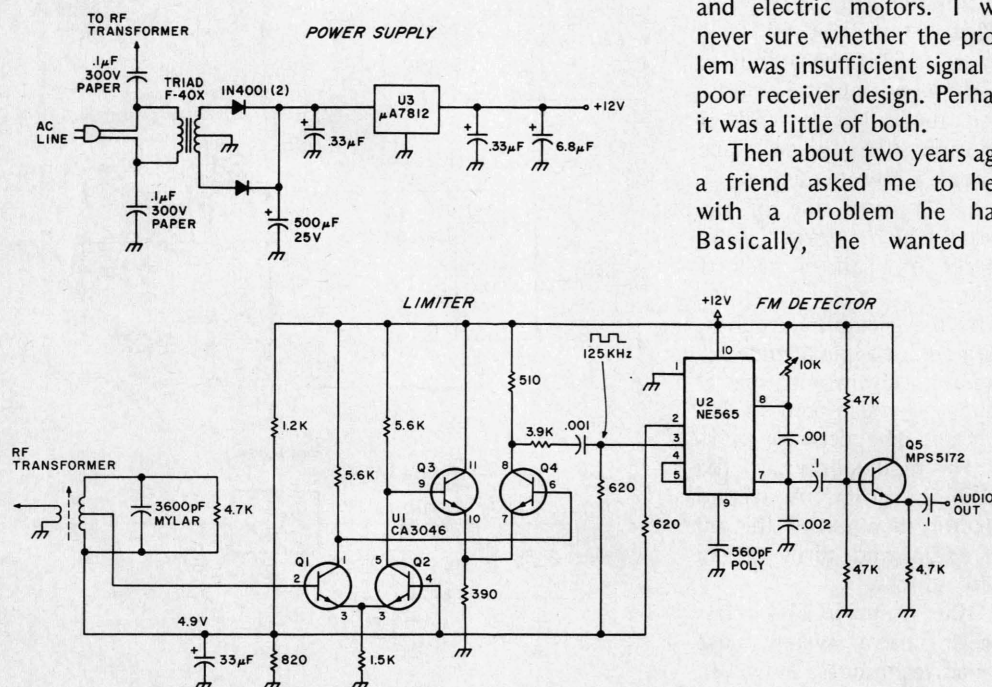


Fig. 4. Rf front end. Notes: See notes (1), (2), and (3) in Fig. 2. (4) Q1, Q2, Q3, and Q4 are part of U1.

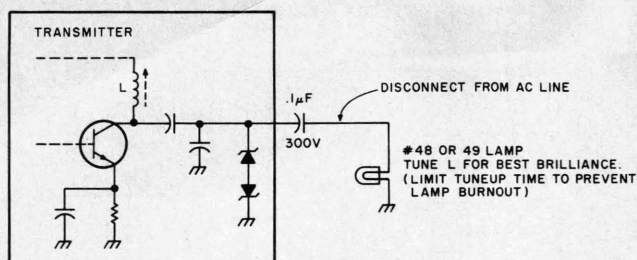


Fig. 7. Remote transmitter tuning setup.

transistor array, U1. The limited signal is then reduced in amplitude by a resistive divider and sent to U2, a phase locked loop FM demodulator. The resultant audio is then coupled to Q5, an emitter follower, which provides a low output impedance.

Next the signal goes to the two tone decoders, U3 and U4. U3 is tuned to 1400 Hz to recognize the A tone, and U4 operates on the 1700 Hz B tone. Presence of the correct tone causes pin 8 of the appropriate decoder to go to ground. If U3 is activated, its output goes low, turning off Q6, whose collector then goes toward 12 volts. When C1 charges sufficiently, SCR Q7 is turned on. Conduction of Q8 turns on the LED to show that an "on" signal has been received. At the same time, the relay is energized, closing its contacts which are used to control an external load. When the A tone disappears, Q7 remains latched on.

Tone B causes U4's output to go low, turning off Q8, which turns on Q9. When on, Q9 shunts current around Q7, turning it off. Then, when tone B disappears, Q9 turns off, causing the LED to go out and the relay to de-energize.

Bits and Pieces

That's most of the system, but a few more comments and suggestions may help you

with your WW system. There are several special components used in this project that helped make it successful. The power transformers were obtained at low cost and serve admirably. In a second version of the system, I used some 15-volt, 1-Ampere transformers and bridge rectifiers. 12-volt transformers or 24-volt center-tapped units could give marginal operation with low line voltage. If you use the lower voltage transformers, just be sure that there is at least a 3-volt dc margin above the 12-volt regulator output, or they will not work.

The transmitter's output coil is rather important. Taking a lead from the National Semiconductor application note, I tried re-winding an i-f can to use here. It worked, but results were poor, due to the high dc resistance of the winding. The coil I used was of unknown origin, and I had to remove turns to make it tune. A J. W. Miller type 23A685RPC coil should work as well.

The receiver's input coil is an i-f can modified as per a modified version of the National Semiconductor application note printed in the *Popular Electronics* article, "A Wireless Audio System for Remote Speakers." Here, both the transmitter and receiver coils were obtained by rewinding miniature transistor radio i-f cans. Fig. 6 shows the required

windings. Although the original article called for Toko transformers, I found that the Radio Shack 273-1383 worked just as well. To perform the modification, first carefully pull the "guts" of the can out the bottom of the shield. Now unscrew the black ferrite cup from the plastic threaded base. Next, cut off the wires to the solder terminals and remove the fixed ceramic capacitor. You will find that the small bobbin containing the windings is held in place only by beeswax. A little heat and careful prying removes it to strip the old wire off and add the new windings. Be careful not to forget which wire is which while you rewind, or you'll have to redo the job as I did (the first time). When soldering the windings onto the base pins, be careful with your iron. Too much heat will destroy the plastic base.

Tune-up of the transmitter is fairly simple. The only frequency adjustment is for the rf output. A frequency counter is recommended. The transmitter output stage is best tuned up by using a no. 48 or 49 pilot lamp as a dummy load. Connect the transmitter output stage as in Fig. 7, and tune the output inductor for maximum brilliance. A few crude tests of rf injected into the line showed this method to be a good compromise between simplicity and maximum signal.

The tone oscillators have no tuning adjustment, since their exact frequencies are not important. To be safe, check the two output frequencies and be sure that they are at least 15 to 20% apart. If they are too close, the receiver's decoders cannot distinguish between them. Change either the 110k or 91k resistors, should any readjustment be needed.

To tune the receiver, a frequency counter and oscilloscope are suggested. First set up a signal source (the matching transmitter is best) to send a weak signal into the receiver. Connect the scope

across the tuned portion of the transformer and tune for maximum signal. To adjust U2, U3, and U4, first short circuit their inputs to ground through a 0.1 µF capacitor, then set their vcOs on frequency. Hook up a counter to pin 4 of U2 to read the vco or to pin 5 of U3 and U4 for their oscillators.

In some cases, I found that some low-impedance loads on the power line, particularly high-wattage incandescent light bulbs, caused severe signal attenuation. Usually this doesn't happen, but for long-distance transmission — 200 or 300 feet — lamps close to the transmitter load the signal heavily. To eliminate the difficulty, I placed a choke in series with the offending load. Commercially-available chokes are awfully expensive, but Fig. 8 shows an adequate home brew version.

Parting Comments

A high-performance WW remote control system has been described. Duplication for similar applications should be simple for experienced radio amateurs.

The same ideas can also be adapted for other uses. For example, if long distances aren't involved, the National Semiconductor transmitter circuit can be used with the tone encoders I've described. Or perhaps the remote speaker idea can be duplicated using my power amplifier. In fact, I've been toying with using my remote speaker to monitor my ham receiver from anywhere in the house. Still another possibility is to make a fancy wireless intercom using circuits from both sources.

WW is a very handy transmission technique. I hope that perhaps I've eliminated some of the mystique involved so that other radio amateurs and experimenters can use it for their own purposes. My thanks go to Gary Kirchner WA3YES for the impetus to take on this project. ■

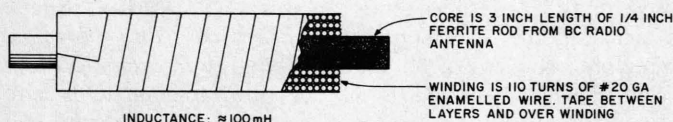


Fig. 8. Rf choke to minimize signal loss in incandescent light.

The schematic diagram illustrates a tuning meter circuit. It begins with an **AUDIO INPUT** connected to a 741 op-amp configured as a voltage follower. The op-amp's non-inverting input (pin 3) is biased at 0V using a 1K resistor to ground. The output (pin 6) is connected to a 0.1μF capacitor, which then feeds into the input of an NE565 timer. The NE565 is configured as a buffer, with its output (pin 2) connected to the base of a 2N5655 transistor. The transistor's emitter is grounded, and its collector is connected to the **METER ADJUST** terminal. The meter is a **TUNING METER** with a range of 0-1mA and is zero-centered. The circuit is powered by a +12V supply and a -12V supply. Various resistors and capacitors are used for biasing, timing, and signal conditioning throughout the stages.

73 Magazine has published a series of interesting articles on facsimile during the last several years: "An IC Facsi-

mile Receiving Converter," Brockman, January, 1974; Ralph Taggart's series, "A Satellite FAX System You

Can Build," August, September, and October, 1975; "A Fast Scan Facsimile System with SSTV Compatibility," Winkler, March, 1973; and "Facsimile for the Radio Amateur," Dean, August and September, 1971. I suggest that this material be explored before getting into the project. In particular, the article by Bob Dean (of Alden Electronic & Impulse Recording Equipment Co.) should be reviewed. It is an informative and comprehensive report on amateur facsimile.

There is one definite prerequisite before getting underway. A little Western Union Telefax transceiver will be needed for a source of gears and bearings (and some electronic components). These "deskfax" units are still plentiful on the surplus market. If there is any difficulty in finding one locally, I can supply a source (\$5.00 for the unit plus shipping). Please include a self-addressed

Build a FAX from Scratch

-- repeater control, with ID

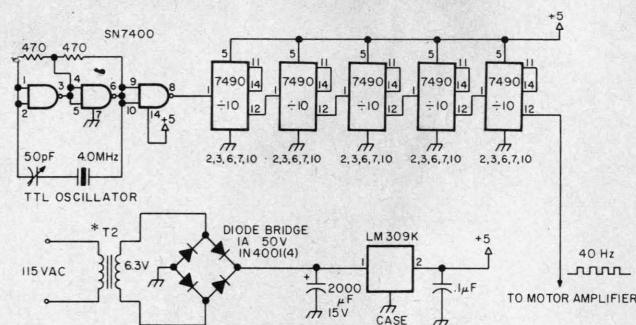


Fig. 2. Precision ac source. *Filament transformer from deskfax.

stamped envelope with your request.

Electronics

The demodulator, Fig. 1, is merely a RTTY terminal unit. A suitable lead resistor and diode have been added to optimize it for facsimile. This circuit, employing an NE565 phase locked loop, was presented by Nat Stinnette W4AYV in the February, 1975, issue of *Ham Radio Magazine*.¹ The demodulator works on the FM principle. The voltage-controlled oscillator of the NE565 locks onto the incoming signal and tracks it between the frequencies of the black (1500 Hz) and white (2300 Hz) tones, resulting in a corresponding dc shift at the output. A voltage comparator connected to the output of the PLL produces plus and minus voltages to activate printing transistor Q1. A micro-center milliammeter functions as a tuning indicator.

The precision ac source, Fig. 2, supplies 40 Hz to a 50 W class B motor amplifier, using a pair of 6550 tubes connected as zero-biased push-pull. Inasmuch as the deskfax operated at 180 rpm (60 Hz supply), 40 Hz is required to control the speed of the synchronous drive motor to 120 rpm. The 120 rpm rate is required to rotate the weather map and some other photo transmissions. The TTL crystal oscillator is a "surefire" one from *ET*, February, 1974,

p. 34. When others are cranky, this one will work!

I used the following procedure to select T_x , the motor amplifier output transformer (see Fig. 3). The method requires a known composition resistor, VTVM, and filament transformer (part of a more extensive routine by W0MKF in *Ham Notebook* for determining values of unknown inductances). The setup will only give ball park values of impedance, but that's close enough:

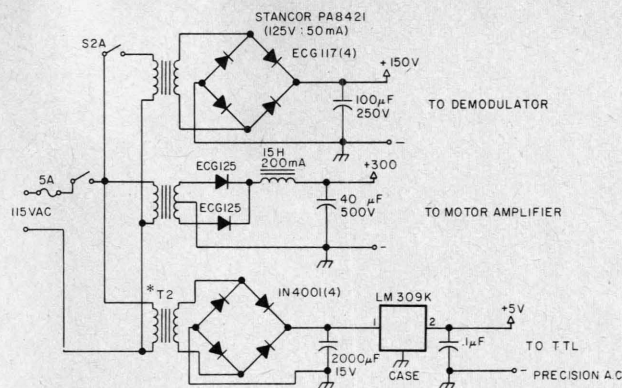
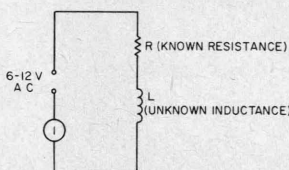


Fig. 3. Power supplies. *T2 — 6.3 V from deskfax. S2 — DPST switch.

1. Measure ac voltage across known series resistor.
2. Compute I from E/R_L .
3. Measure ac voltage across inductance.
4. Compute Z from E/I .

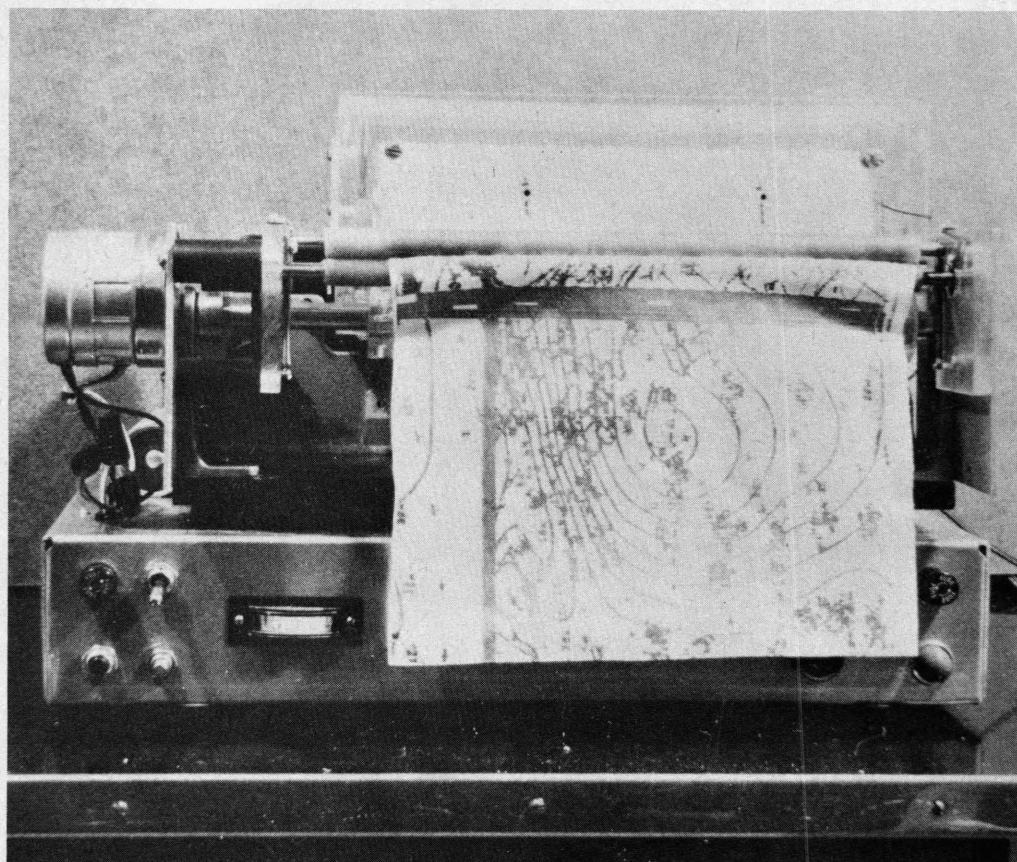
Applying the method, in turn, to deskfax motor windings, primary and secondary of a surplus (unmarked) power transformer on hand, the following values were obtained:

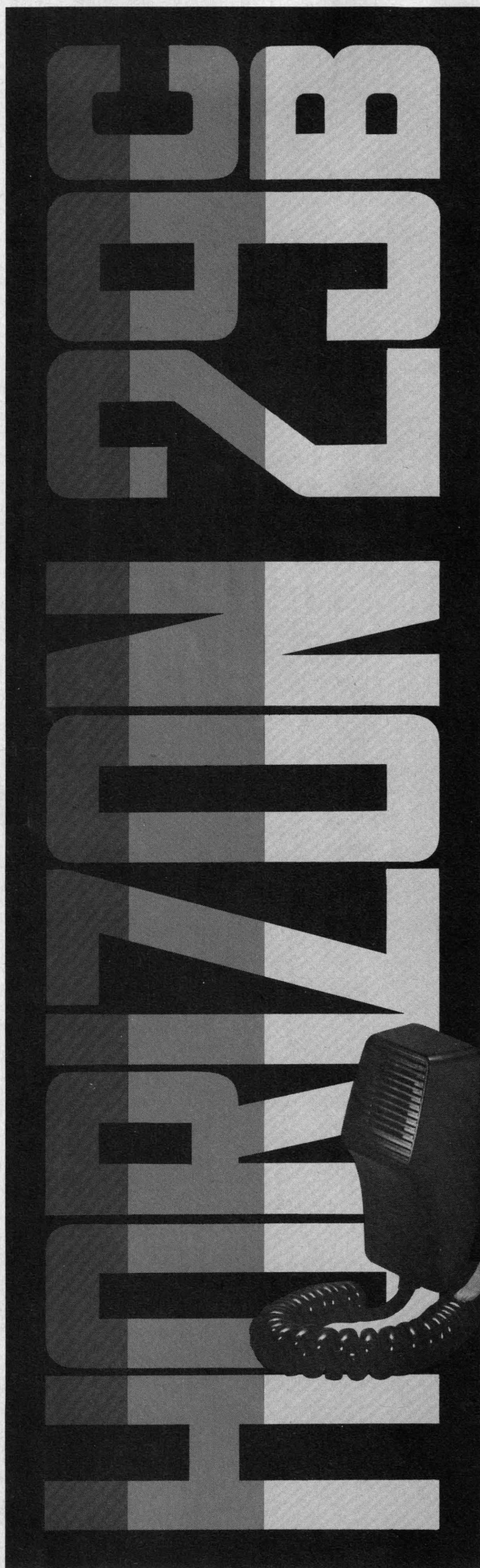


$$\text{Deskfax: } \frac{Z \text{ (Ohms)}}{750}$$

Primary: 1400

Secondary: 20k





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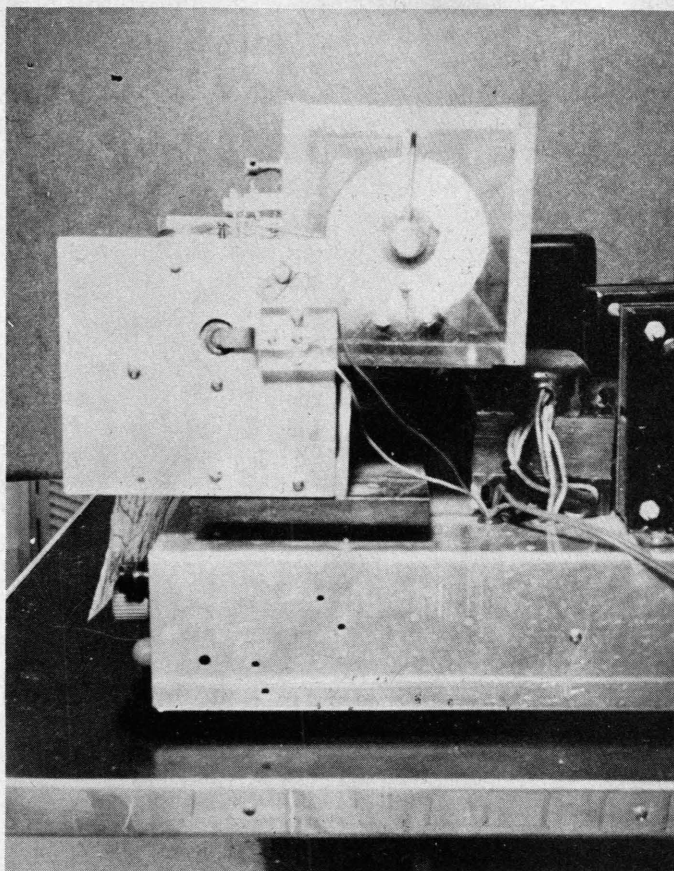
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Helix contactor.

That's the best combination my junk box could produce — anyway, it works!

In the motor amplifier, Fig. 4, L and C (7 Henry choke and 2 mF capacitor) filter the square wave 40 Hz from the TTL source to an acceptable sine wave for the amplifier. In selecting the output transformer, T_X , select a husky one, 115 V primary, with a secondary around 350-0-350 V at 175-200 mA.

Mechanics

As a first step, dismantle the deskfax. Disengage the reversing mechanism. Rotate the drum and slide it slowly to the left. As you look through the hole in the drum, you will see a brass screw in the shaft. Remove the screw. Loosen the collars and the bearing. Now the half inch shaft can be removed from the left. Remove the bearing and strip it. Remove the following electronic com-

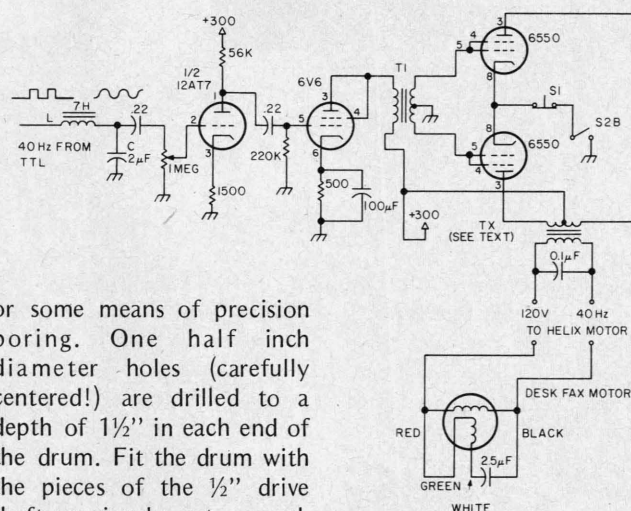
ponents: the push-button switch, the 6.3 V filament transformer, and the 2.5 mF motor capacitor.

Cut the cradle in two, making the cut 4-3/4" from the left end (the bronze sleeve bearings should be covered to protect them from filings). Next, cut the 1/2" steel drive shaft as follows: from the left end of the shaft, 1 piece 4-7/8"; from the right end of the shaft, measuring from the shoulder, 1 piece 2" long. (This allows 1/2" clearance between the end of the helix drum and the right bearing.) See Fig. 5. Center and drill and tap the bearing end of the 2" piece for a 1/4" 4/40 screw (this will be a wiping contact).

Helix

I've used Wink W6WMI's system in constructing the helix. (See Fig. 5.) The helix drum is a plexiglas rod 1 1/2" diameter, 10" long. This is a step that will require a lathe

Fig. 4. Motor amplifier. T_X — see text. T_1 — Stancor A4713. C — Stancor C1707. S1 — N.C. SPST push-button switch. All resistors 1/2 Watt 10%.



or some means of precision boring. One half inch diameter holes (carefully centered!) are drilled to a depth of 1 1/2" in each end of the drum. Fit the drum with the pieces of the 1/2" drive shaft previously cut, according to Fig. 5. Drill and tap for a 4/40 screw 1/4" in from the right end of the drum. This screw must go through the wall of the drum and make good electrical contact with the drive shaft. Drill and tap for 6/32, 1/4" screws to anchor the #22 nichrome helix wire. The screws are located 1/4" in from each end of the drum and positioned to form a 370° lead (one spiral turn occupying 9" with about 5° overlap at each end).

Looking at the right end of the drum, rotation will be counterclockwise. The nichrome wire is to be wound clockwise (sweep will be from right to left as contact is made with the blade). Snug the wire up as tightly as possible on the drum, then pass the free right end over to be fastened by the contacting screw.

Cradle

Using a 3/16" bit, drill any additional holes needed in the base of each of the two pieces of cradle previously cut. Set the pieces on a plexiglas sheet 3/8" by 4 1/2" by 14 1/4". Temporarily install the helix and align it so that it turns smoothly. Mark the mounting holes, then drill and tap for #8 screws. A 13" by 17" by 3" chassis is used for assembling the recorder. Place the

plexiglas base (above) on the chassis. At this time install whatever you intend to use for shock mounts. There should be at least 4. (Excellent shock mounts are available from the deskfax, but there are only 3.)

Paper Humidor

An aluminum plate 1/8" by 3 1/2" by 10" is to be installed on the flange of the right cradle member (this flange formerly supported the optical system). Position the plate so that it overlaps the forward edge of the flange 1/2" and is aligned with the right end of the cradle. Before mounting the plate, using a 3/16" bit, drill holes in each corner of the plate for mounting the paper humidor.

The Alfax paper must be kept moist for the electrolytic process to work; so while the plexiglas enclosure does add to the expense of the project, it is an essential part of the recorder. IPS Weldon #3 solvent (or other suitable brand) is used in assembly. You'll need an applicator. I paid the additional price and had the distributor cut the pieces to size. See Fig. 6. The box goes together easily — just tack the members with a few pieces of Scotch tape, apply the sol-

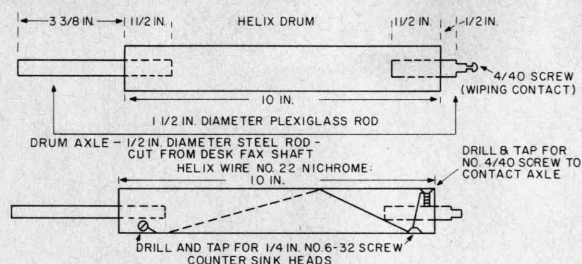


Fig. 5. Helix.

vent, and in a matter of seconds the pieces are *really* welded. Note that paper will be fed through a 1/8" slit, formed by pieces A and B of the front of the humidor.* Next cut out two supports for the Alfax paper. Use 3/8" plexiglas and cut to the dimensions of Fig. 6. Center these at the ends of the interior of the humidor and cement with solvent. Align the top, then drill and tap for a 1/2 inch 6/32 screw at each end. Place a roll of Alfax paper in the humidor and replace the top. Drill and tap for a 6/32 screw 2" long that will extend down through the cover and contact the hub of the paper roll. These should be about 9/16" in from the end of the cover and in line with the center of the hub. Do this for both ends. The 2" 6/32 screws installed will function as "brakes" to control the flow of paper.

Blade Assembly

Cut a piece of 3/8" plexiglas to 1 1/4 by 9 1/2". Install steel pins at each end, as in Fig. 7. (Material for the pins can be taken from the desk-fax push-button switch assembly.) Cut a piece of 1/2" aluminum angle and install with 4/40 screws as shown. The two screws at each end at the top will anchor two phosphor-bronze motor brush springs. The three screws along the forward edge will

secure the blade,² wedging it between the plexiglas and the aluminum angle. (The National Weather Service and some TV stations use Alden facsimile recorders that use a blade in the form of a steel loop. These are discarded after a little wear and are a good source of material for the blade required here. 9 1/2" sections can be cut from the loop.)

Paper Feed Assembly

See Figs. 8 and 9. The paper feed mechanism is driven by a 5 Watt, 115 V, 1/2 rpm, 60 Hz synchronous motor. This is a Hurst, Model CA, available directly from the Hurst Mfg. Co., Motor Division, Princeton, Indiana, currently at \$18.06. This speed (1/2 rpm) with 1/2" rollers, produces a 9" by 11" recording. Not exactly symmetrical, if one is a purist. The rollers are made by slipping 10-3/4" lengths of rubber tubing (1/2" O.D., 1/4" I.D.) over 1/4" steel rod (the material should be turned and ground shaft or drill rod). The driven roller is 13-1/8" long; the idler is 11-7/8". Ball bearings (NICE type 1602

DC) are used for the driven roller. The bearings are 11/16" O.D., 1/4" I.D. The tension springs, E and F, are 1-3/4" long, cut from the traversing mechanism (spring) of the deskfax. Reference Fig. 9: the 3/16" diameter holes at O and P are to be at a depth of 3/16" on the inner surfaces of supports A and D. These will retain a 3/16" rod (preferably brass) to smooth and align the paper as it emerges from the slit in the humidor. The bottom of the rod should align with the top of the helix and be positioned about 5/8" forward from the center line of the helix. (This detail is not shown in the exploded view of Fig. 8.) Two aluminum "end plates" of 1/8" material, each 4 1/2" by 5 1/4", support the paper feed assembly and the contactor for the helix wire. See Fig. 10. A coupling 1" long will be required to couple the 3/16" motor shaft to the 1/4" roller rod (you'll probably have to make this). Place the plexiglas supports, A and D, on their respective end plates, and align the top and forward edges. Mark the mounting holes and the center line of the driven roller shaft. Drill a 5/16" hole for the motor shaft. This is to be expanded to 1/2" after the motor mounting holes have been marked and drilled.

A 3/4" diameter hole is to be drilled in the right end plate, 1-3/4" from the top edge, 2-7/8" from the forward edge. This is for the rotating contact (a 4/40 screw) in the right end of the

helix drive shaft. The contactor is merely a piece of "springy" brass mounted on a small piece of 3/8" plexiglas.

Getting It All Together

Install the fiber gear and then the 7/16" collar at the left end of the helix drive shaft. Install the helix drive motor in the left cradle member. Place the helix in the sleeve bearings and mount the assembly to the plexiglas base. Position the aluminum end plates with the bottom edges aligned with the bottom of the cradle, and with the rear edges aligned with the rear edge of motor support. Attach these to the ends of the cradle with #8 screws (2 in the lower edge and 1 on the uprights that house the sleeve bearings). Press the ball bearings into the 11/16" holes in the plexiglas supports (A and D). Slip the paper smoothing rod into each support, seating it at O and P (Fig. 9). Install the driven roller. Attach this assembly to the aluminum end plates. The left support, A, will require 3 spacers 1 1/2" long. These can be made up of 3 sets of 3/8" O.D. spacers for #8 screws, 1/2" long (H. H. Smith #2119) and 1" long (H. H. Smith #2117). Slip the coupling onto the left end of the roller shaft; mount the motor. Mount the springs and adjustable bearings (Fig. 9) for the idler roller. Install the aluminum base plate for the paper humidor. Place the humidor on the plate centered with the helix, with its front 1-1/8" from the center line of the helix. Mount and attach the humidor to the plate (use 1/4" screws). Slip the left pin of the

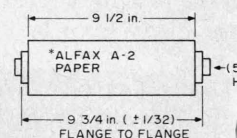
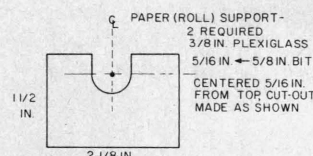
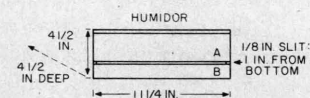


Fig. 6. Humidor and roll supports. Humidor (3/8" plexiglas): 2 pcs. 3-3/4" x 3-3/4" (ends); 1 pc. 4-1/8" x 11-1/4" (back); 1 pc. 3-3/4" x 11-1/4" (bottom); 1 pc. 4-1/2" x 11-1/4" (top); 1 pc. 2-5/8" x 11-1/4" (front); 1 pc. 1-3/8" x 11-1/4" (front). Assemble with IPS Weldon #3 solvent.

Fig. 6(a). *Alfax plexiglas roll supports with collapsible core.

*Plexiglas, solvent, and rubber tubing were obtained from General Rubber & Supply Co., 3118 Preston Highway, Louisville KY.

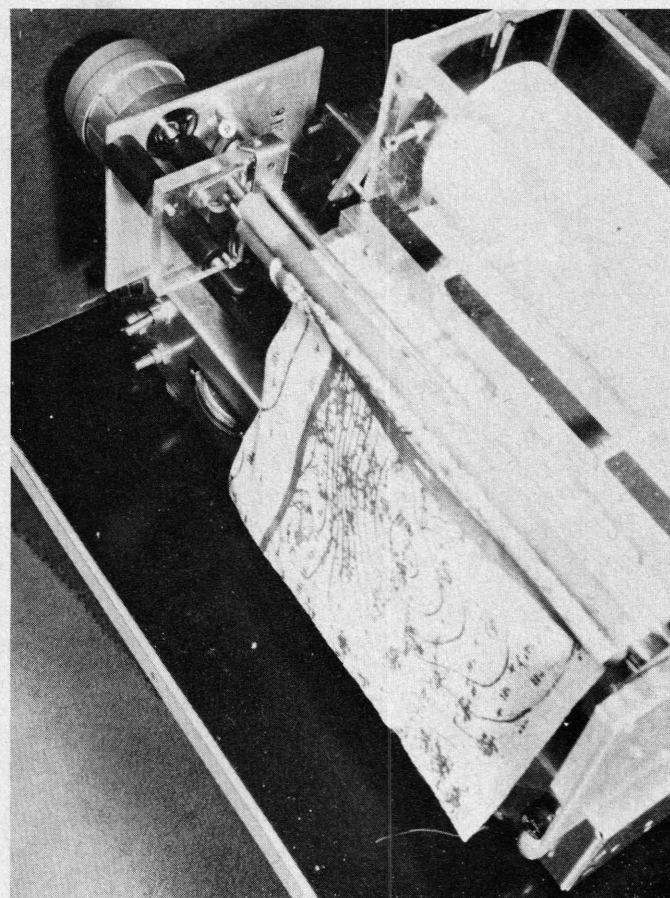
assembly into its mounting bracket (a piece of 1/8" brass, 1/2" by 3"). Rest the blade on the helix, center it, and, with the blade assembly leveled, mark the right end plate for a 1/8" bearing hole. Similarly mark and mount the left bracket. (Small thin pieces of brass, attached with 4/40 screws to the outer surfaces of the bearing members, will prevent lateral movement of the blade assembly.) Two #8 screws, 1 1/4" long, support the phosphor-bronze motor springs (Fig. 7). Use a #55 (or similar) bit and drill a hole adjacent to the head of each screw. Drill and tap the front wall of the humidor for these #8 screws, 1" from the top and 1 1/4" inch from each end. Two 1/4" hex spacers, 1" long, threaded for a 6/32 screw (H. H. Smith #2324) are drilled and tapped for a 4/40 screw at 1/4" from an end. Run a locking nut down on two 4/40 screws, 3/4" long, and then install in the hex spacers. Solder compression-type springs (I cut them from a transistor heat sink clip) to the tips of each 4/40 screw. Thread small gauge copper wire through the holes in the #8 screws. After installing them in the front wall of the humidor, secure the ends of the phosphor-bronze springs (the other end of the springs should be fastened to the minimum angle of the blade assembly). Solder a lead to the right spring. Attach the

helix contact to the right end plate, centering it on the 4/40 screw head from the drive shaft. Put a few drops of #10 oil on the sleeve bearings and worm gear. Turn the helix by hand to see that everything is still aligned. Set the collar and fiber gear in place. There should be just a tiny amount of backlash with the fiber gear and worm — and a similar "tiny" amount of end play between the collar and left bearing. Apply power to the helix drive motor to make any final adjustments of the 4 mounting screws necessary for smooth operation. Pour a small quantity of water into the humidor, and, assuming all of the electronics are functioning perfectly(!), place a roll of paper in the recorder and look for NSS!

Operation

Fleet facsimile broadcasts are made from NSS, Washington, D.C., on the following frequencies (kHz): 3357, 4975, 8080, 10865, 16410, and 20015. Incidentally, publication H.O. 118, available from the U.S. Naval Oceanographic Office, Washington, D.C., contains a wealth of data on worldwide FAX weather broadcasts (as well as RTTY). There is also information on interpreting weather codes.

This is the procedure I used with a Drake SSB receiver, an R4B, to adjust the PLL demodulator. Tune in the NSS signal (upper side-



Paper feed detail.

band). Most transmissions are 2300 Hz (white). Set R1 of of about 10 minutes duration, with 800 Hz shift between 1500 Hz (black) and Tuning to the 2300 Hz tone

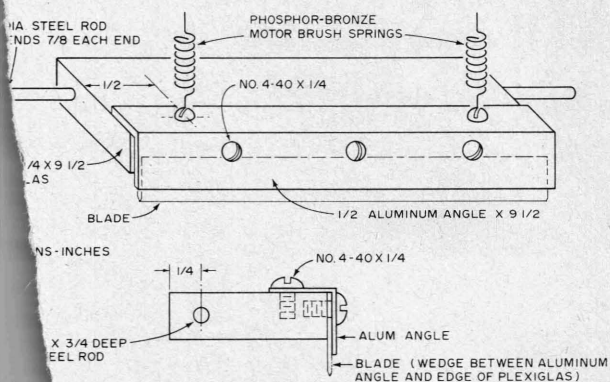


Fig. 7. Blade assembly.

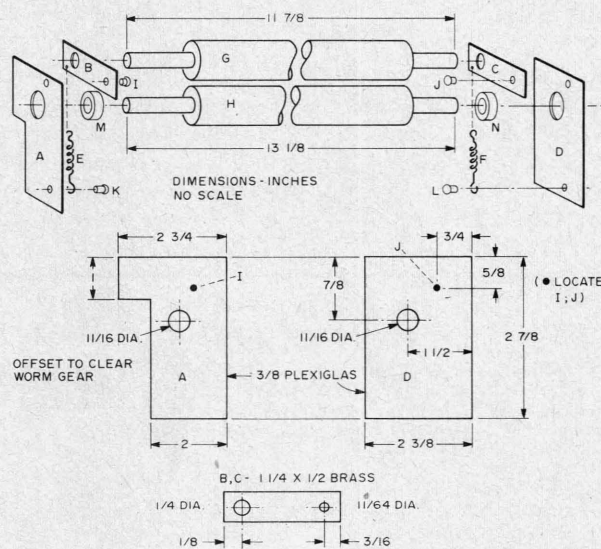


Fig. 8. Paper feed assembly (not to scale). A, D: left and right supports, 3/8" plexiglas. B, C: top roller bearing, adjustable brass, 1/8" x 1/2" x 1-1/4". E, F: tension springs. G, H: rollers. I, J: #8 screws, 1/4". K, L: 4/40 1/2". M, N: ball bearings.

should cause the zero-center milliammeter (tuning indicator) to deflect toward maximum negative. The 2300 Hz tone is transmitted between maps, so this is a good time to check the tuning indicator and set it for maximum deflection. Remain tuned to the 2300 Hz tone until a start tone is heard, indicating the beginning of the next map transmission. This will be a lower tone (I believe it is 675 Hz). This will be transmitted for 10 seconds, and will be followed by about 30 seconds of phasing pulses. During this time, with the recorder running, you should see the recording coming in as a field of black with a white line (about 1/2" wide) in the field. Momentarily activate the phasing switch (push-button) and note the shift of the white line to the right. Continue to pulse the phasing switch until you have the

white line at the left edge of the paper. You now have your drum phased with that of NSS for proper reception of the map. The copy will probably look thin and skimpy with this setting of R1. Back off on R1 slightly and retune the signal. Note that you are able to control the width of lines in the recording (to some degree) by the setting of R1.

The recorder will copy wirephoto transmissions made at 120 rpm. Wirephoto transmissions will be on lower sideband (and inasmuch as press frequencies are private, it would be inappropriate to publish them here). I can say that occasionally I've tuned above 15690 kHz and come upon a signal that sounds like a flock of nervous wild turkeys, and with a little jockeying around the lower sideband, was able to get a picture.

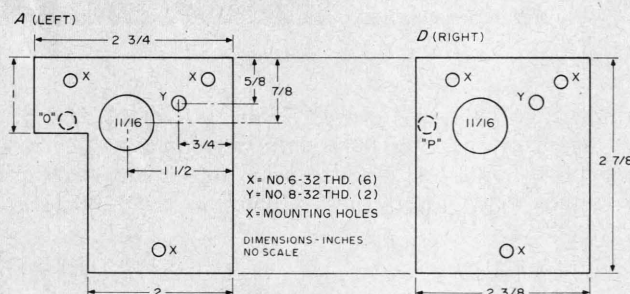


Fig. 9. Paper feed supports (3/8" plexiglas). Drill and tap each piece for 8/32 1/4" screw, 3/4" from forward edge, 5/8" from top. Drill 11/16" hole for ball bearing, on each piece 1 1/2" from forward edge, 7/8" from top. O and P: 3/16" diameter holes to a depth of 3/16" on inner surface of each piece, 7/8" from top, 2 1/4" from forward edge, to retain 3/16" smoothing and aligning rod (brass, 3/16" x 12 1/4").

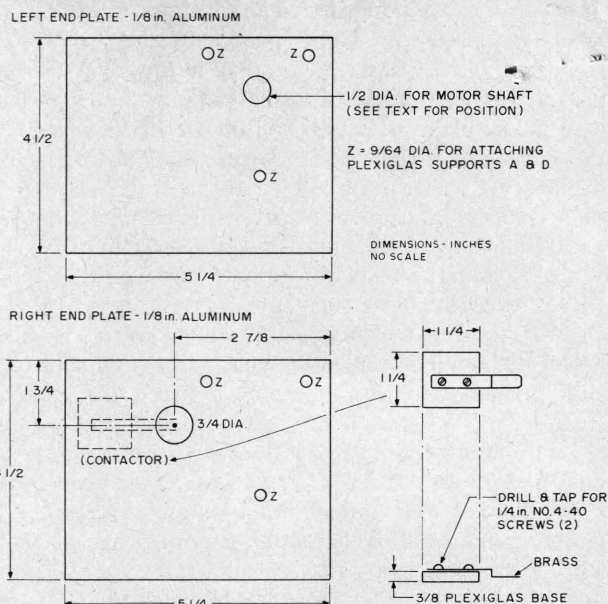


Fig. 10. End plates for paper feed assembly.

Options

If a smaller format is preferred, Alfax paper is available in various widths: 8 1/2, 6-1/8, 5 1/2 inches (1/4 inch is to be added to accommodate flange thickness). A 19 inch format with paper feed of 96 lines per inch is more or less standard in facsimile transmissions. Lines per inch are increased with a smaller format, according to the relation:

$$LPI = \frac{\text{Helix rate (rpm)}}{\text{Paper feed (in./min.)}}$$

One would merely have to rewind the helix, select a paper feed motor with slower rpm (or decrease the diameter of the rollers), and adjust the paper roll supports accordingly.

Acknowledgement

Credit should be given to Lindsay Winkler W6WMI for much of the mechanical detail here. As an inspection of his "Fast Scan FAX System" (March, 1973, 73 Magazine) will show, I've drawn freely on his ingenious design of helix and paper feed mechanisms. I am indebted to "Wink" and also to N6 Stinnette W4AYV for simplifying the project. ■

References

1. PC boards for this circuit available from Nat Stinnette W4AYV, P.O. Box 1043, Tavares, FL 32778. Undrilled board \$4.75; \$25.95 for a wired and tested unit.
2. Paper, helix wire, and blade (steel loop) material can be obtained from Alden Co., Westfield, MA 01581.

Type	Description	Price
11C01FC	High Speed Dual 5-4 Input	OR/NOR \$15.40
11C05DC	1 GHZ Counter Divide By 4	\$74.35
11C05DM	1 GHZ Counter Divide By 4	\$110.50
11C06DC	UHF Prescaler 750 MHz Flip/Flop	D Type \$12.30
11C24DC	Dual TTL VCM	\$2.60
11C44DC	Phase Freq. Detector	\$2.60
11C58DC	ECL VCM	\$4.53
11C70DC	600 MHz Flip/Flop With Reset	\$12.30
11C83DC	1 GHZ 248/256 Prescaler	\$29.90
11C90DC	650 MHz ECL/TTL Prescaler	\$16.00
11C90DM	650 MHz ECL/TTL Prescaler	\$24.60
11C91DC	650 MHz ECL/TTL Prescaler	\$16.00
11C91DM	650 MHz ECL/TTL Prescaler	\$24.60
95H90DC	250 MHz Prescaler	\$9.50
95H90DM	250 MHz Prescaler	\$16.55
95H91DC	250 MHz Prescaler	\$9.50
95H91DM	250 MHz Prescaler	\$16.50

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NEW Fairchild VHF Prescaler Chips

TUBES	New		New	
	IP21	519.95	6146A	\$4.25
	2E26	\$4.00	6146B/8298A	\$5.50
	4X150C	\$18.00	6360	\$5.50
	4X150A	\$15.00	6661	\$1.00
	4X250B	\$24.00	6680	\$1.00
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TV/VCR Tuner Applications

Part Two of three—you may want to keep Part One, from last month's 73, handy.

Hugh Wells W6WTU
1411 18th Street
Manhattan Beach CA 90266-4025

Voltages developed for use with tuners must be regulated and heavily filtered to reduce noise and voltage variations. Noise and voltage variations can cause tuner frequency instability, since tuners are voltage-operated devices. Regulators such as the 7812 and LM317 work well in tuner power supplies.

The mixer and band control circuits operate on +12 volts. A pot attached to the +12 volt bus is used to provide an adjustable voltage of approximately +7 volts which is connected to the AGC terminal. The actual AGC voltage value is unique to the tuner and must be determined experimentally while listening to a signal. In operation, the AGC voltage may be varied with a pot and used as an RF gain control.

Tuner tuning is accomplished by providing a voltage which is variable from 0 to 29 volts. The 29-volt supply may remain fixed, but the voltage applied to the tuning pin of the tuner can be varied with a pot. The most satisfactory tuning resolution is obtained by using a ten-turn pot. The addition of a vernier knob to drive the pot is helpful. Even with vernier adjustment, the tuning can seem a little erratic. To smooth out the tuning action, a 1- μ F capacitor may be connected from the tuning pin to ground. Although a larger capacitor value may be used, the voltage time constant could thus be made too long, allowing frequency adjustment overshoot should the tuning knob be moved too rapidly.

If the tuner has an AFT terminal, you can utilize the feature for fine tuning. If unused, the terminal must be grounded. To be used for fine tuning, the terminal may be connected to the wiper of a 5k pot. Voltage for the pot should be a low value not exceeding 2-3 volts. One end of the pot can be connected to a positive voltage and the other to ground or to -1 volt.

Construction

Fig. 1 shows the typical wiring and voltage requirements to support a TV/VCR tuner used as a frequency converter.

Mechanical construction is not critical, allowing a lot of freedom for the builder to choose the mounting of the tuner, pots, switch, etc.

Although the regulators are operating well below their dissipation rating, I suggest that they be mounted against a PC board or onto a small heat sink. The objective is to reduce the tendency for a voltage shift as a function of heat rise.

Dial calibration

After the converter is operating and the IF established at a specific frequency, a "look-up" table may be constructed to

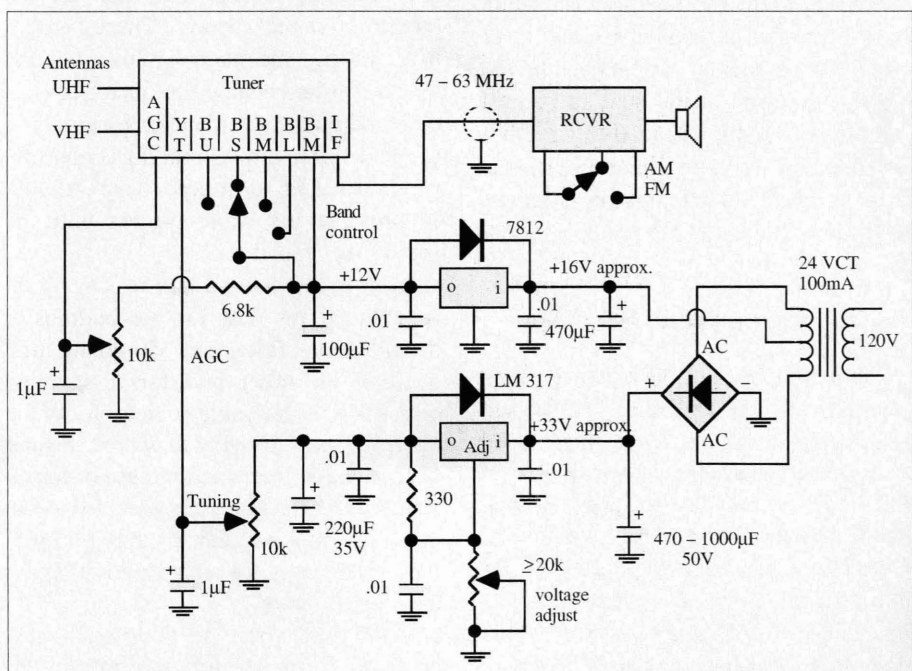


Fig. 1. TV/VCR tuner as a wideband frequency converter.

The Atlantic Telegraph Company of Great Britain

Cyrus had spent a good deal of his own money and a lot of his friends' getting this far, and his search for more "sports" in the US was being hampered by his critics. Undaunted, Field set off for England to gain additional financial support, and by the end of 1856, he had secured British backing of more than \$2 million (a lot of money for the times) and organized the Atlantic Telegraph Company of Great Britain.

Back in business, Cyrus and his original friends (The Newfoundland and London Telegraph bunch) immediately set out to pick up where they had left off. By August, 1857, he pushed off from Ireland, laying cable toward Newfoundland. After four days and 350 miles, the cable snapped in two and half a million dollars went down the drain.

At this point, it was obvious that he needed a new cable-laying method. The scheme these guys came up with, though, was unique. They decided the best way was to use two ships, starting in mid-Atlantic, one headed for Ireland, and the other, Newfoundland, each paying out cable en route. Cyrus got it all going by mid-1858 with the two ships and all (he had to convince his backers this would work).

After several highly embarrassing experiences, both ships returned to Ireland to regroup. This was an occasion probably only Field the undaunted would have been able to handle. He had to convince his backers, and indeed he did. Underway again at the start of August, 1858, the two ships headed toward their destinations. Without major event, each ship reached its destination after six days. The Atlantic cable was in. But was it?

A hero's welcome turns to rebuke

In the first week of operation the cable worked well, but by the end of the second week of August, 1858, trouble began. In the meantime, celebrations were the order of the day for the "Atlantic Cable Heroes," the type of activities Cyrus Field was adept at promoting. The culmination of this hype was to be the exchange of messages between President Buchanan and Queen Victoria on August 16th, but by the 16th the line was

almost dead. In spite of this, the messages got through (after two days of trying).

While the line continued to work to some degree well into October of 1858, word was spreading about the problem. The ensuing outrage that followed from both sides of the Atlantic resulted in Field's lying low for a few months.

After the dust had settled, Field and his backers decided to blame the problem on "faulty cable manufacture" which led to the premature failure of the cable (the "dust" had included allegations of fraud and the like). No one really knew what actually happened, but one thing they all knew was that they were broke, and new money was hard to get. Weeks ran into months and on into years with Cyrus out beating the bushes, trying to raise more money.

Field went on without success for over five years, both in England and the United States. The Civil War was raging throughout America at the time, and it, on its own, generated some interest in the need for a cable. But even the British at the time were reluctant to spend more money on the deal, having been recently burned by another undersea cable fiasco (in the Red Sea).

The Anglo-American Company

In 1860, halfway through the project, Cyrus W. Field had been forced into bankruptcy. He had consumed most of his personal fortune and that of his creditors on the venture. There came a time, during the nadir of his financial woes, that he became the object of ridicule and public avoidance. Nonetheless, he had persevered, making countless trips back and forth across the Atlantic to promote his cable project until he finally succeeded.

After seven years, Field was again on speaking terms with his old buddies at the Atlantic Telegraph Company and most of the other investors associated with his Atlantic cable project(s). As the Civil War was drawing to a close, he and his colleagues recruited the use of a passenger ship, the *Great Eastern*, for use in cable-laying for a year. By July of 1865, they were ready for another go at laying an Atlantic cable.

In the usual Cyrus W. Field style, they departed from Ireland after much fanfare. One thing he never learned was to

avoid publicity, but with all the experience this bunch had, they *had* learned to continuously test the cable while en route. They were also using the latest in undersea cable design which, unfortunately for Field, still had several years of maturing to go. As cable defects were detected, they reeled the cable back until each bad section was located; they then repaired the defect, and continued on.

With all going well and about two-thirds of the distance traveled, they somehow lost the cable. Ten days later, they found it... only to lose it again, this time for good—or so it seemed. Cyrus must have personally known the coiner of "Try, try again," because off he went to London again, to reassure his backers that he knew what he was doing—and he did it; again, he organized a new company (the Anglo-American Company); he reloaded the *Great Eastern*; and headed for Newfoundland, although delayed almost a full year.

"Day after day, night after night, her great pistons glided in their channels, and the cable poured over the stern without a hitch. At last they had learned how it was done!" —Harlow, *Old Wires and New Waves*.

On July 27, 1866, the *Great Eastern* entered Trinity Bay, Newfoundland, still in communication with their terminus at Valentia Bay, Ireland. The cable was good. The cable was in!

Upon securing the July 27, 1866, cable, Cyrus turned the *Great Eastern* back to sea to look for his lost 1865 cable. Armed with better grappling equipment than he'd had the previous year, he found the 1865 cable on September 5, 1866, spliced it, and brought it ashore at Trinity Bay on September 8. This second cable worked perfectly, and the world now had *two* Atlantic cables. (Oddly, this stunt did more to restore Field's credibility than the docking of the July 27 cable.)

After losing an estimated \$10 million over the preceding twelve years, Field eventually paid back all of his backers and investors, made money for everyone associated with the venture, and was then able to retire as a great world hero. By 1875, using the technology Cyrus W. Field pioneered, virtually every populated continent of the world was linked via cable. 73

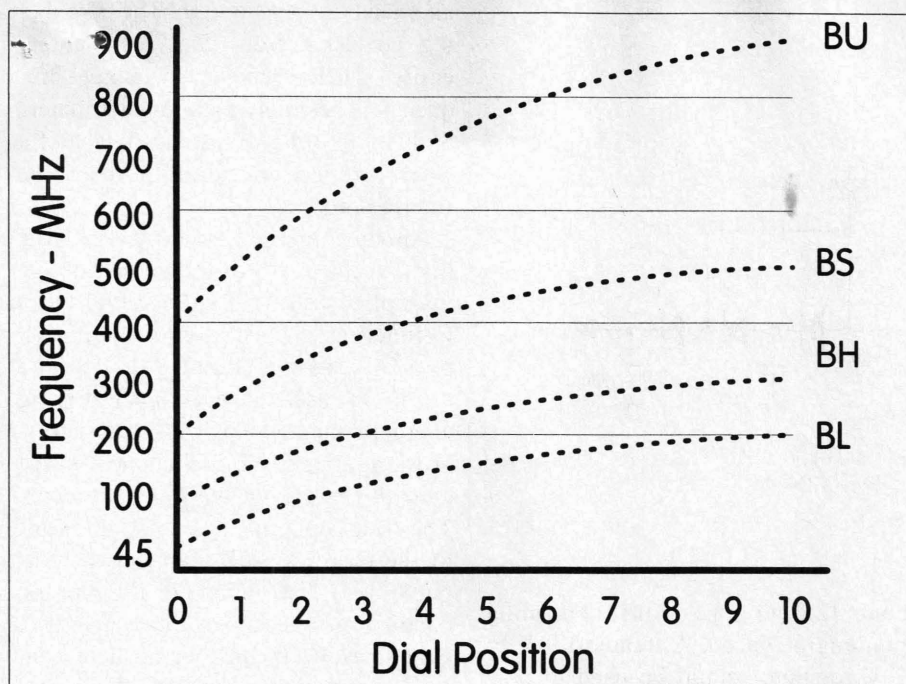


Fig. 2. Dial-to-frequency translation.

translate dial position markings to frequency. Specific frequency settings are then interpolated between known frequency calibration points.

Another method is to construct a calibration curve as shown in Fig. 2. Although the tuning curves are not linear, a straight line may be drawn between known frequency checkpoints. The readability resolution of the chart is set by the spacing between the X and Y coordinates on each axis.

Instead of using dial position markings on one of the chart's axes, you may use the tuning voltage for greater accuracy of frequency repeatability. TV/VCR tuners are quite reliable in their frequency resetability as a function of the tuning voltage. The use of a DVM to measure the tuning voltage to the 1st

and/or 2nd decimal place is desirable.

If a signal generator is not available for determining frequency calibration points, known spot frequencies may be identified and logged. Signals such as cordless phone, FM BC, aircraft, public

"These tuners have one major drawback in mobile applications."

service, taxi cabs, ham bands, weather, TV audio channels and cellular phone can provide the necessary frequency references for rough calibration. In most cases it is sufficient to identify the band edges for each service.

Take advantage of tuner availability—build a wideband frequency converter and tune in to the action taking place in the 45-900 MHz spectrum!

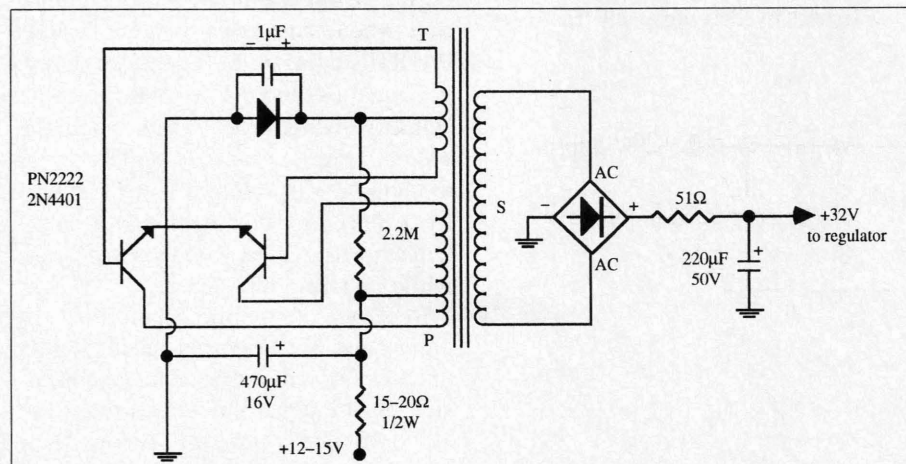


Fig. 3. DC-DC switching converter. All diodes are 1N4148. Operating frequency is approximately 56KHz. Transformer core: powdered iron, OD-0.4 to 0.5in., ID-0.125 to 0.150in. Windings: PRI-15T #26 enamel wound bifilar, SEC-30T #30 enamel, TIC-1T #30 enamel wound bifilar.

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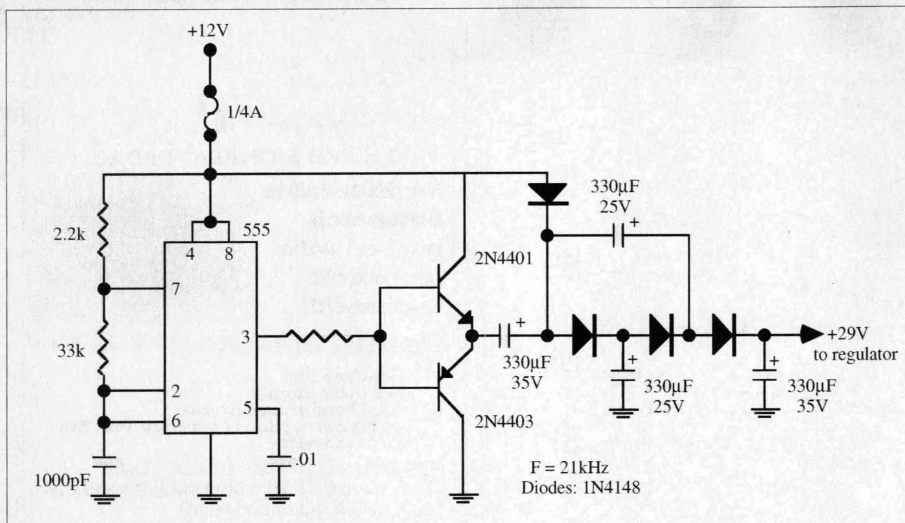


Fig. 4. 555 driving a voltage multiplier.

Mobile power

Because of their mechanical stability, TV/VCR tuners work very well in mobile applications. However, there is one major drawback that requires attention. TV/VCR tuners are voltage-operated devices and are not tolerant of changing voltage values. Therefore, mobile operation presents a real challenge for the experimenter. With semiconductors available, the challenge can be minimized.

The voltage requirements of a tuner are +12 volts and an ultra stable tuning voltage supply of approximately +30 volts. The +12 volts should be regulated because it is used to power the mixer/osc, band control and AGC circuits. Although most tuners prefer +12 volts, they will function without noticeable degradation down to about +10 volts. An automobile's electrical system is a suitable source and has a voltage value which can vary from

about 12.6 to 14.7 volts depending upon engine speed. A standard 9 V to 12 V regulator could be used to provide stable voltage for tuner use except during the periodic loss in regulator headroom at perhaps a 12.6 V value when the battery terminal voltage falls. A simple zener diode regulator is normally sufficient and solves the problem nicely for the tuner.

The real challenge for mobile operation is the +30 volt requirement for the tuning voltage. There are several ways to generate a high voltage from a 12-volt supply. Two of the ways, a DC-DC converter and a voltage multiplier, are presented here. Perhaps the easiest method of obtaining the desired voltage is to build a DC-DC converter designed around a powdered iron core as shown in Fig. 3. Most any powdered iron core, except ferrite, will work in this application. Switching is accomplished by driving each transistor

alternately into saturation. The core, being powdered iron, does not saturate easily. Therefore, the switching frequency is desirably high (approximately 56 kHz) in order to take advantage of a small core size and to aid in the ripple filtering process.

Another method of obtaining the high voltage is to utilize a voltage multiplier. But for a voltage multiplier to function, it must be driven with AC or a DC chopper circuit. Fig. 4 shows a 555 IC used as a chopper driving a complementary pair of transistors. Working alone, the 555 cannot source enough power, but the transistors can. The transistors are driven from cutoff to saturation by the square wave output of the 555. There is little or no conduction overlap in the transistors, which results in little or no heat generation—therefore, a heat sink is not required.

With the AC voltage component generated by the chopper, the voltage multiplier will increase the output voltage to approximately 29 volts (which is marginal for tuning the high frequency end). It is difficult to obtain a higher voltage from the circuit shown, because only one voltage rail of the multiplier is moving, not two as would be the case with an AC source. Single rail switching limits the function to a maximum voltage multiple of three, minus losses.

Fig. 5 shows an LM317 regulator that is required to regulate the output of the +30 volt source. Adjustment of the regulator is critical in order to obtain the highest possible stable voltage. The correct setting is found during a battery non-charging condition when the source voltage is low. Pot R1 is adjusted to reduce the output voltage just enough to obtain output stability, which will occur when the voltage drop across the regulator rises to about 4 volts. Output voltage stability achieved by this method will be retained as the battery voltage rises/varies while under normal charge.

Because of mobile operation, all parts must be supported mechanically to prevent vibration damage. Heat-sinking, though, is not required since all parts are operating below their dissipation factor.

Next time: Spectrum analyzer, signal source, multifunction signal device. **73**

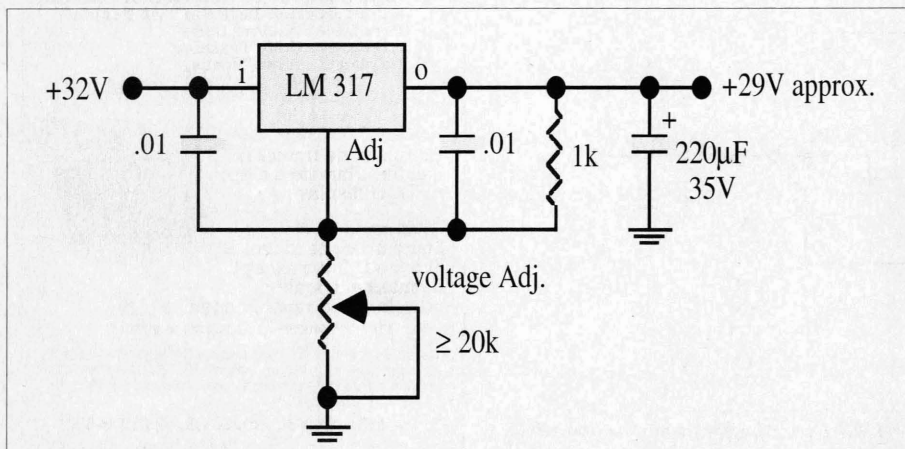


Fig. 5. 29-volt regulator.

TV/VCR Tuner Applications (Part 3)

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Spectrum analyzer

Perhaps one of the handiest tools for electronics is the spectrum analyzer. In essence, it is a receiver having an oscilloscope attached which is used to provide a spectral view of a band of selected frequencies. The amount of spectrum to be viewed can be varied for either a close-up view of the signal, or a greater amount for viewing all of the adjacent signals within the band. The usefulness of the spectrum analyzer is great, with many applications. To name a few: transmitter sideband analysis, spur generation analysis, frequency spotting and identification, identifying modulation mode, harmonic signal analysis, and use as a wide frequency range receiver.

The spectrum analyzer described here is constructed using relatively easy to obtain components so that it is a project within reach of any ham experimenter.

The two main components are the TV/VCR tuner and a receiver. Although a receiver board may be constructed, almost any receiver capable of tuning to the tuner's IF (46–63 MHz) and having an AM detector will create a fine spectrum analyzer. I assume that you prefer to use a pre-built or commercial receiver, so a simple home-brew receiver will not be discussed. Many suitable receivers are available that will tune the 46–63 MHz region, such as six-meter AM and 30–50 MHz WWII military receivers. Even a 27-MHz CB AM radio used in conjunction with a 47- to 27-MHz converter will work well. The narrow IF bandwidth of the receiver provides narrow dispersion for viewing signal carrier and sideband characteristics.

The block diagram in Fig. 1 shows the major components and interconnects for a spectrum analyzer. When in operation, a sawtooth voltage is applied to the tuning voltage line, causing the tuner to

sweep across a band of frequencies. Any received signal within that band will be translated to a DC voltage whose amplitude is relative to the strength of the signal which is then displayed on the vertical plane of the oscilloscope. In order to maintain a steady display on the screen, the horizontal scope sweep must be synchronized with the sawtooth generator. The sweep speed is set just high enough to eliminate viewing flicker. A high sweep speed does nothing to improve the display. Any basic oscilloscope may be used, but perhaps the main requirement for the scope is to have a DC input so that it will respond to the rectified output signal from the receiver's AM detector.

Fig. 2 shows the complete schematic for the spectrum analyzer with the exception of the receiver details. To drive the tuner sweep circuit, a unijunction transistor was chosen to generate the sawtooth waveform, but any sawtooth generator circuit will work in this application. In order to provide a fairly linear sawtooth ramp, a JFET is used as a constant current source for charging the timing capacitor. A voltage pulse is taken from the B2 terminal of the UJT to provide a sync pulse for the scope.

As an alternative sweep source, some oscilloscopes provide a horizontal sweep sawtooth output which may be divided down with resistors and applied to the sweep width pot. The sweep voltage amplitude as generated within the scope may be as high as 100 volts, but the maximum desired sweep voltage is about 10 volts when applied to the sweep width pot. On the average only 0.5 to 1 volt is required to provide a normal spectral display (sweep width).

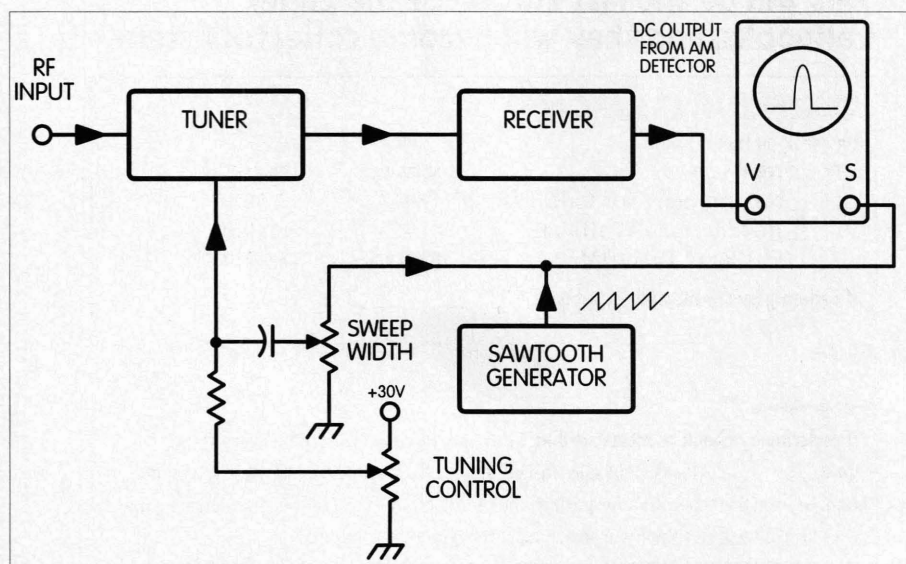
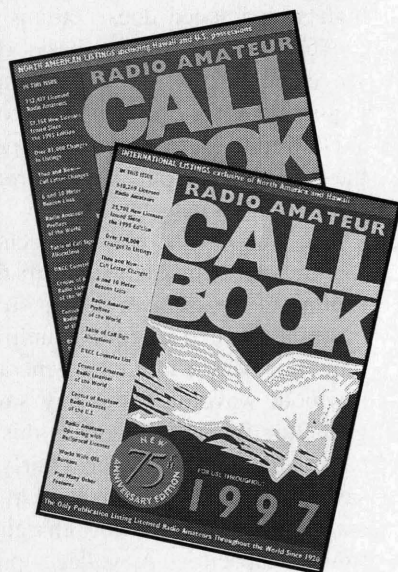


Fig. 1. Spectrum analyzer block diagram.

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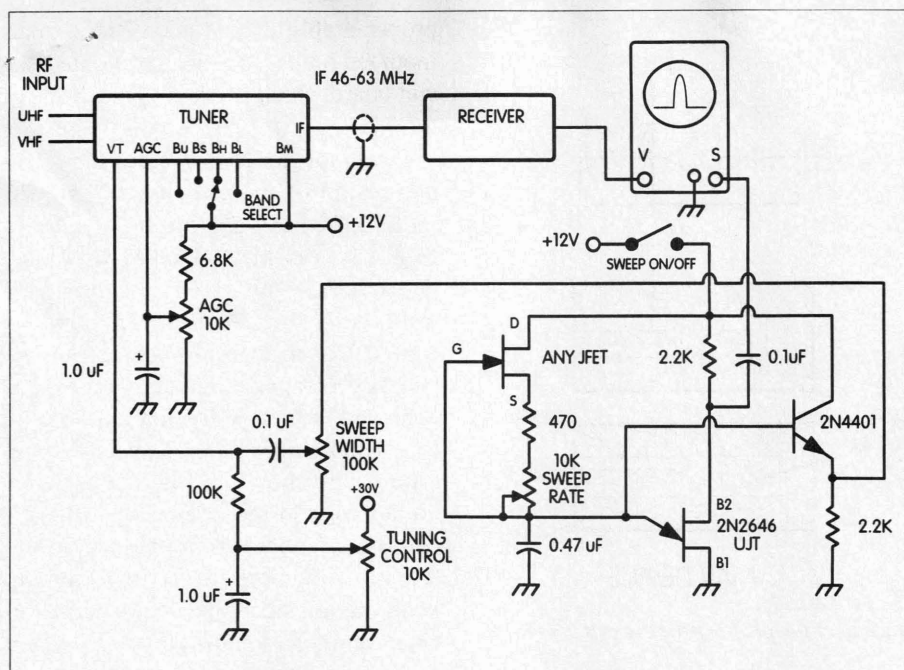


Fig. 2. Spectrum analyzer schematic.

Power supply regulation, although not as critical for the +12 volt line, is recommended to reduce tuner drift. Regulation is very critical for the tuning voltage circuit, however, and care must be taken to remove all ripple and noise. The tuning is performed by using a 10k pot across the +30 volt source. For vernier tuning control, a 10-turn pot and a counter knob are recommended.

Almost any small 24-28 volt, 100-200 mA, power transformer is suitable for use in the spectrum analyzer. It is suggested that filter capacitors have a value equal to or greater than 3300 μ F to keep power supply ripple and noise to a minimum.

Mechanically, the circuit is very stable and can be mounted in any convenient manner to suit your taste.

Signal source

TV/VCR tuners were designed to be the front end of a receiver. As such,

they contain a local oscillator which is used to mix with the incoming signal to create an IF of approximately 47 MHz. The tuner, as a receiver front end, covers a received frequency range of 45 to 900 MHz. The oscillator operates above the received frequency and is offset by the IF (46-63 MHz). Therefore, the oscillator operates over the range of approximately 90 to 950 MHz. The actual oscillator range is dependent upon the specific tuner. Typically, TV oscillators tune from 91 to 940 MHz while VCR oscillators tune from 108 to 955 MHz.

Some tuners are set up with a phono connector near the oscillator circuit that outputs the oscillator signal to the PLL frequency control circuit. In most cases, the output is frequency-divided prior to exiting the tuner case, making the output generally unusable. However, some tuners provide a direct oscillator output, and when that is the case, the tuner will work as a signal generator without modification. When

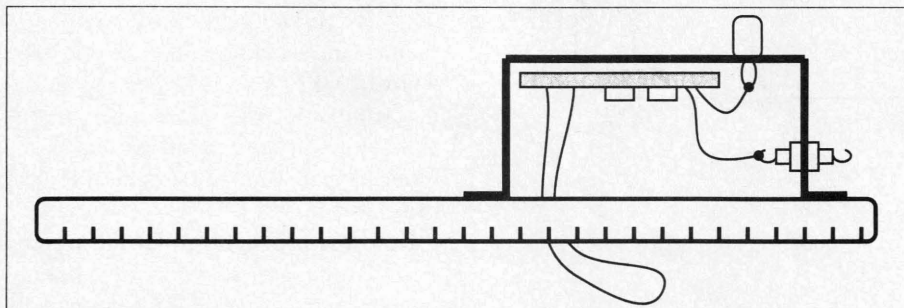


Fig. 3. Amplifier mounting and coupling. The pickup loop feeds signal from the oscillator to the amplifier board.

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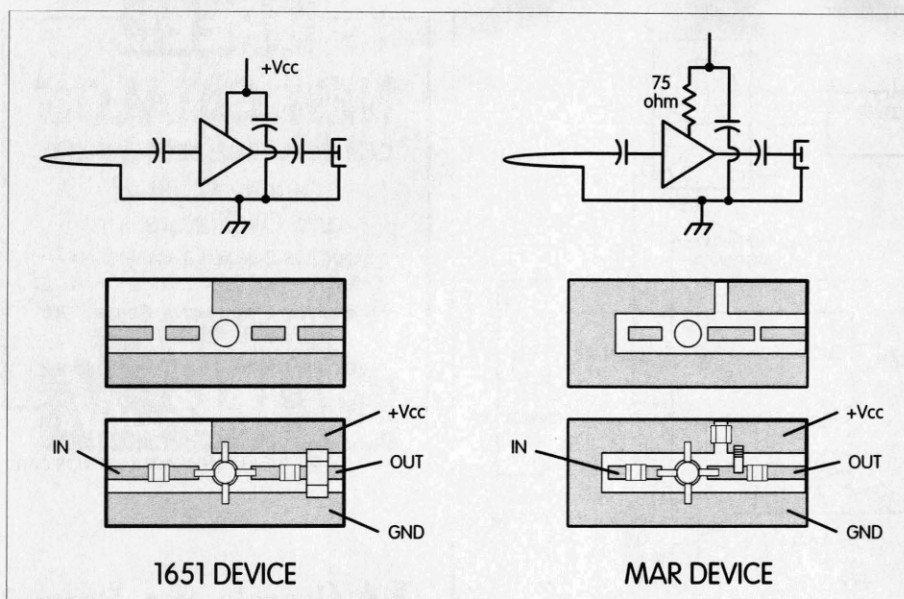


Fig. 4. Schematic, PC board layout, and parts placement for the amplifier board, shown for the two devices discussed in the article.

access to the oscillator is not available, the tuner may be modified slightly to obtain a direct signal from the oscillator. There are many techniques available to capture the oscillator signal, but only one technique will be described here.

The concept for capturing the oscillator signal is shown in Fig. 3 and involves drilling a hole in the tuner cover to provide access for a wire pickup loop to pass through the hole, allowing the loop to lie close to the oscillator circuit. The loop picks up a signal which is fed into a broadband amplifier. A coupling capacitor is connected between the loop and the amplifier to satisfy the isolated DC input requirement of the amplifier. The amplifier provides a constant load on the oscillator and is used to amplify the signal captured by the loop. If the

tuner cavity is large enough, the amplifier may be mounted inside of the tuner rather than on the cover. Otherwise, the amplifier must be mounted outside of the tuner. To prevent radiation from the amplifier, the circuit must be enclosed in a metal box made from a tin can. The box is made RF-tight and soldered to the tuner cover. The box may be of any convenient dimensions, but it must be RF-tight.

Power for the amplifier is in the range of 4.5 to 5 volts. To prevent chip heating, the voltage at the chip should not exceed 5.5 volts. The power to the amplifier must be fed via a high capacity feedthrough.

A phono connector may be mounted on the amplifier box and used as an RF output connector. The amplifier output impedance is 50 ohms. To provide

proper amplifier operation, the chip is mounted on a small piece of printed circuit board which is glued to the inside of the box.

The amplifier chip used in the original design of the circuit was an NEC UPC1651G. As an alternate part, a Mini-Circuits MAR-1 or MAR-2 may be used. A 75 ohm 1/4 watt resistor is used with the MAR series chip to source Vcc to the chip. An output impedance of approximately 40 ohms is achieved, which will source a 50 ohm line satisfactorily. The respective chip circuits are shown in Fig. 4. Capacitor values used in the circuit are not critical, as any value from about 0.001 to 0.5 uF will work. It is important to keep the physical size of the capacitor small and lead lengths kept very short. Chip capacitors are preferred when available.

After making the tuner into a signal generator, it will still function as a normal signal converter/tuner. The addition of a pickup loop will shift the oscillator frequency a small amount, but not enough to prevent the tuner from functioning as designed.

Multifunction signal device

If you haven't noticed, all of the new pieces of ham equipment abound with a multitude of functions and capability. As a result, some marvelous equipment and innovations are offered to the ham community. Equipment designed and built by hams may also exhibit multiple innovations when a little thought is applied.

Parts/components for ham-designed equipment come from some surprising sources and places. Therefore, it should come as no surprise to find TV/VCR tuners designed into a multifunction device having the following functions: frequency converter, spectrum analyzer, signal generator, sweep generator, and standard deviation signal source. All of the indicated capability is achieved with external circuits and with a minor modification to the tuner to extract a signal from the local oscillator.

Before we conclude, let me remind you that the actual frequency band obtained from a tuner is determined by the individual manufacturer's design. Therefore, all frequencies indicated

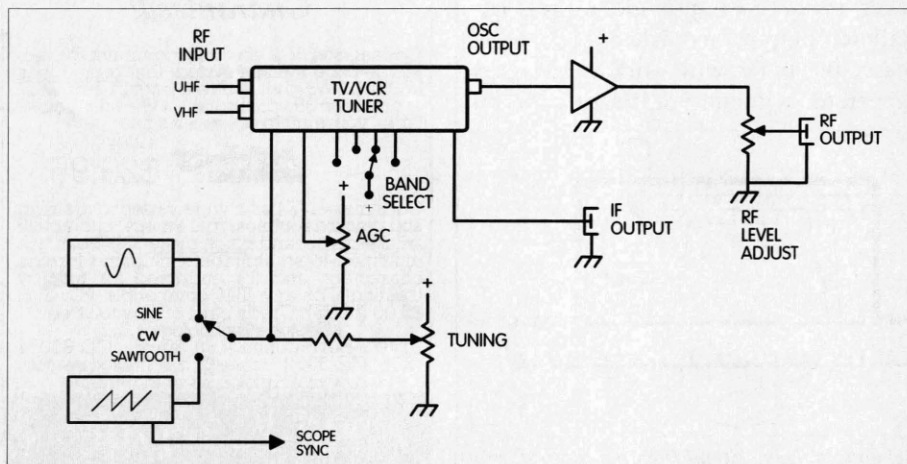


Fig. 5. Multifunction signal device block diagram.

here can only be used for reference. Refer to the previous articles in this series for more information.

Tuners typically require two source voltages for proper operation. Positive 12 volts is applied to the mixer/oscillator and band control circuits, and is divided with a pot for the AGC. The AGC may be used as the RF gain control. A variable voltage from 0 to 30 volts is used for tuning. Both of the voltage sources should be regulated to achieve best tuner stability. It is also recommended that large value filter capacitors be used to keep power supply ripple and noise to a minimum.

For a TV/VCR tuner to be useful as a signal generator, a signal output must be obtained from the local oscillator. Some tuners have a phono connector mounted next to the oscillator circuit which was used to output a signal to the PLL. Some tuners have the oscillator output divided before it exits the

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tuner which reduces the tuner's desirability. However, modifying the tuner will allow obtaining a direct signal from the oscillator.

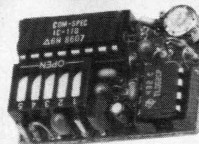
Fig. 5 shows a block diagram of the multifunction signal device with a tuner as the major component. Output from the oscillator is sampled and amplified using a monolithic RF amplifier chip. A pot in the amplifier's output circuit is used to control the signal output amplitude. As a signal generator, three modulation functions are provided. Obviously, when in the CW mode there is no modulation present, but modulation is present when either a sine wave or sawtooth waveform is used. FM is the resulting modulation mode when either waveform is applied.

Fig. 6 shows the schematic for the multifunction signal device. To build the device, it is recommended that the supporting articles be reviewed to gain additional technical information.

Following is a brief description of each function of the device:

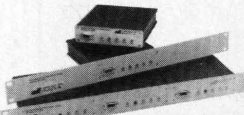
Frequency converter. The TV/VCR tuner was designed to be the front end of a receiver so no tuner modification is required to use it for that purpose.

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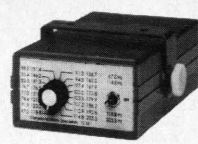


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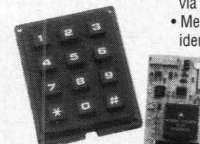
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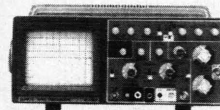
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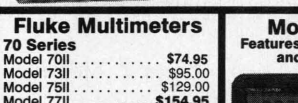


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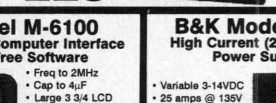
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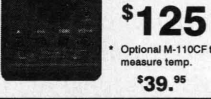
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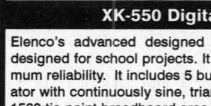
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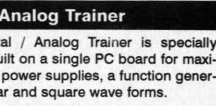
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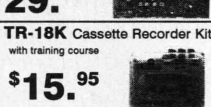


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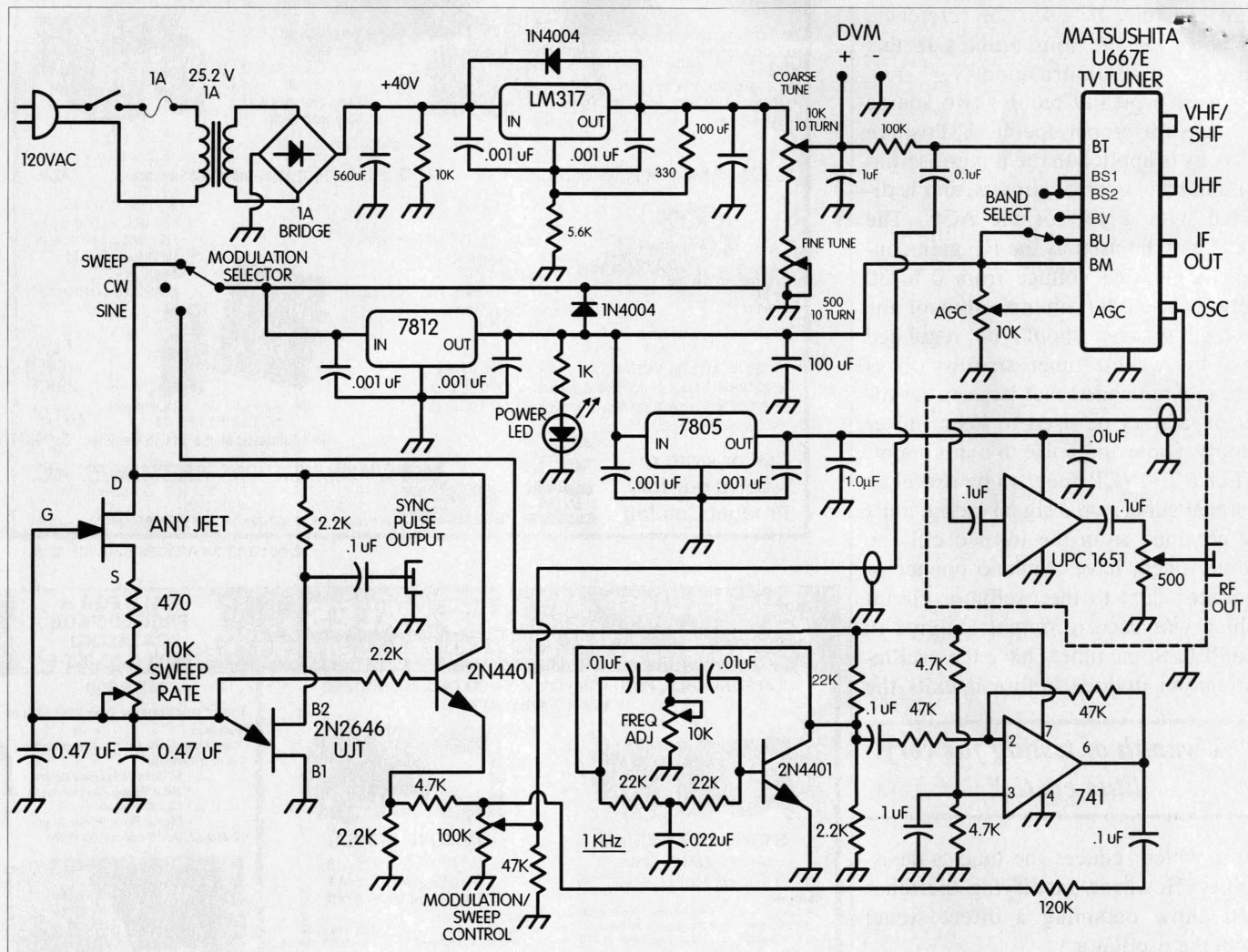


Fig. 6. Multifunction signal device schematic.

Any receiver capable of tuning to the tuner's IF will work well. To name a few: 6-meter receivers, scanners, WWII 50-MHz receivers, and a CB radio used with a 47-27 MHz frequency converter. As a downside, it must be mentioned that tuner drift may be noticed when a narrow passband receiver is used with the tuner. Using the tuner outside of a PLL, the local oscillator frequency stability is subject to voltage and thermal variations.

Spectrum analyzer. The multifunction signal device becomes a spectrum analyzer when used as a frequency converter. An AM receiver with a DC output voltage obtained from the detector is used to provide a signal to the DC input of an oscilloscope. A voltage pulse from the sweep generator is used to synchronize the scope sweep with the tuner's spectrum sweep.

Sweep signal generator. When a sawtooth signal is applied to the generator, the frequency of the generator

will sweep across a band of frequencies and then snap back to the starting frequency. The frequency sweeping process is repetitive and the sweep range is controlled by the amplitude of the sweep voltage.

There are two major objectives when using a sweep generator. The first is to profile the frequency response of a resonant circuit, RF or IF amplifier, or any other narrow passband circuit. The second objective is to provide the necessary sweep voltage for sweeping the tuner over a band of frequencies when the tuner is used as the front end of a spectrum analyzer.

Standard deviated signal source. The objective of modulating the signal generator with a sine wave has two advantages. The first is to provide audio modulation for identifying the output signal when the device is used as a signal generator. The second objective is to generate a signal having a specific value of deviation. In the process of

deviating a signal, sideband energy is created as a function of the frequency of modulation and the rate of deviation. As the amount of deviation is increased from zero, the carrier will pass through a series of null points which are used as a measure of deviation.

A 1-kHz sine wave provides a predictable carrier null vs. deviation at 2.405 kHz at the first null and 5.520 kHz at the second. The carrier null can be observed using a receiver having a BFO centered on the generator's carrier, which will null at a specific deviation as the amplitude of the modulation is increased from zero to that point.

It should be noted that the multifunction signal device can be built and tested a section at a time, or you can build just as much as you need. Whether you build just the spectrum analyzer or the complete multifunction unit, the use of a surplus TV tuner provides a wealth of test capability for very little cost. 73

Inside Digital TV/VCR Tuners

Part 1: How to make them work.

TV tuners have been around for a long time and have really helped hams with numerous projects. The early TV tuners used tuning capacitors and/or variable inductors for channel selection and were later replaced by detent and switchable tuners. Although these tuners worked well in ham projects, the real breakthrough came with the varactor tuners that opened up a whole new world for ham exploration.

The more recent introduction (about 1995) of digital tuners changed the way tuners performed in both TV and VCR applications by increasing tuning stability. Digital tuners are now beginning to find their way to curbside trash bins, where hams are encountering them for perhaps the first time. My introduction to digital tuners began when Pete Haas N8PIF, sent me a tuner and requested information on how to use it. Unfortunately, at the time I knew perhaps less than anyone about them. However, having the tuner in-hand was very compelling, and the information provided in this series on tuners is the result of my study.

While we're on the subject of giving credit to those who have made this

study possible, David Shalita W6MIK followed along very closely with me as the study moved forward. David built up a data test set similar to mine, which will be described later, so that we could work essentially in parallel. Having David work with me was similar to having help pulling up on one's own boot straps. But with help, of course, you gain ground, and we did, as you'll see reported here.

My thanks goes out to both Pete and David for their support in bringing this digital tuner project to an experimenter's beginning. With the digital tuner door opened, only experimentation will tell us how far we might go. If you have an interest in tuners, then please follow along, as this is an exciting technical adventure. It was a tremendous learning experience for me. As you'll see, I'm NOT a digitally oriented person, but I've gained a new perspective on the field.

Digital tuners

There are essentially two types of digital tuners at this point, classed as 3-wire and as 2-wire. Digital tuners, as shown in Fig. 1, look very similar to the more common varactor tuner. I'll

describe the 2-wire first so that we'll be able to spend the rest of the time on the 3-wire tuner.

The 2-wire tuner is the latest entry into the tuner stable, and because of the technology used, it will most likely be around for a long time. Control of the tuner is via separate data and clock lines, and follows an adopted computer protocol (reference Philips Semiconductor's application note AN 036) called I2C, which is a two-way, two-line communication bus between a microprocessor and peripheral ICs. In the case of a 2-wire digital TV/VCR tuner, the internal synthesizer has a unique address, and when addressed by a microprocessor/controller, the synthesizer will acknowledge and then be enabled to receive data via the data bus. From that point on, the tuner provides all of the normal tuner activities such as receiving a signal at a given frequency, and then generates an output signal that is passed along to the rest of the receiver.

The 3-wire bus system used in TV/VCR tuners is the subject of information to be provided in this series. Although the 3-wire tuners are compatible for I2C control via the CLOCK

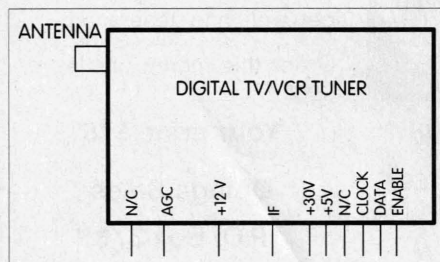


Fig. 1. Typical 3-wire digital tuner is shown. A 2-wire tuner appears to be similar but minus the enable input.

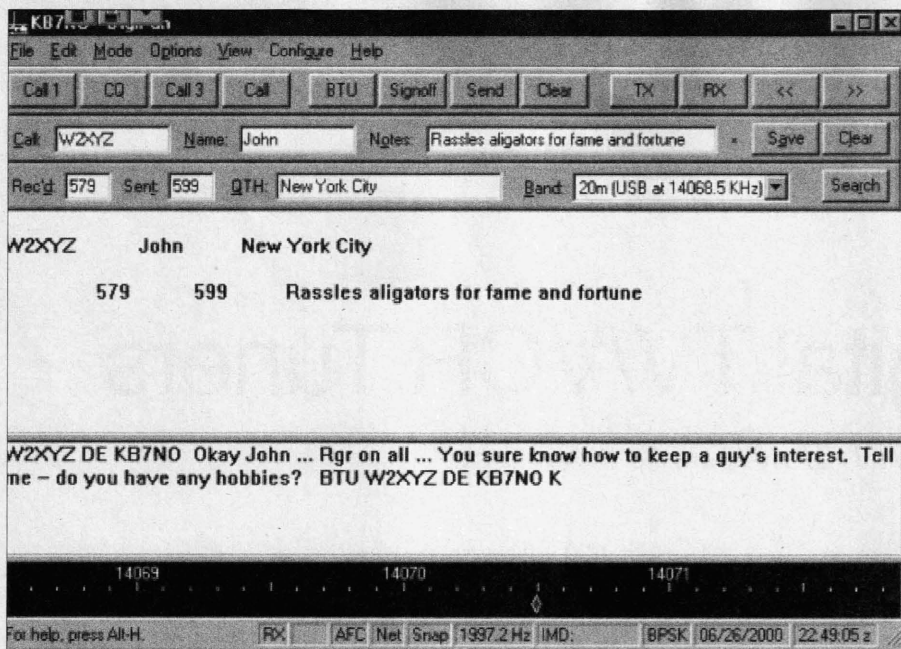


Fig. 1. Screenshot — Here is a shot of the program “everybody is using,” Version 1.2 of DigiPan. The macro buttons at the top can either be clicked with a mouse, or you can use function keys to activate the macro. As a demo, I typed the bits of info in the receive field, then double clicked the callsign which was copied into the “Call” cell. The same was done with the John and the 579. For the other entries, it was necessary to select, click and drag per the help file info. In the transmit field, the two callsign exchanges were entered by using the pre-defined macros. The position of the diamond in the tuning screen at the bottom can show the frequency in use if you tune your radio to the frequency in the “Band” cell at the top. Hidden behind the “always on top” icons (from unrelated software) at the top, is your station callsign; as soon as you configure the free DigiPan software.

The package included the Internet Explorer browser and it was required, otherwise the little 20 to 30 megabyte accounting package would not play. To install the whole shebang required 175 Megs! I rebelled. Actually, the computer wasn't in the correct mode. The drive is partitioned and the major part of the program had to be installed where I had the least room — or else. A call to Microsoft didn't offer a cure, only confirmed the waste of space was part of the process. I decided to scrap the software from the world leader and go buy something from a slightly less egotistical organization. No wonder all the accusations and court proceedings. Hard to believe until you see it first hand. Not much ham software would be sold if programmers used such tactics. There would never have been a PSK31 revolution.

Another look at an intense programming effort for computer control of your transceiver and attendant accessories may offer some interest to you. Some time back, I was made aware of [http://www.trx-manager.com/], which displays a shareware program for control of most popular transceivers. The first time I accessed the URL I happened on the French version and didn't fare well.

The other day I gave it another stab, and there is an English version that the above URL takes you to. If you are curious and would like to see such a program work in your shack, you can download a full working version with a limited trial period. I gave it a try. It worked on my new rig, which indicates they are keeping the software updated. I haven't formed an opinion of its true value to me, but it is worth a look if you have a computer and a transceiver. They give a lot of help and hints for what to do when things don't work, so you should be able to make it do its stuff.

If you have questions or comments about this column, E-mail me [jheller@sierra.net]. I will gladly share what I

know, or find a resource for you. For now, 73, Jack KB7NO. 73

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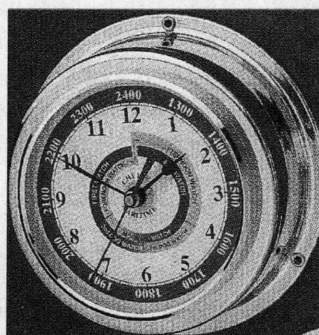
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and DATA bus, the ENABLE bus must be kept separate because the 3-wire tuner does not have a unique internal address.

3-wire data requirements

Those 3-wire tuners that I've encountered to date require that 18/19 bits of data be clocked serially into the internal register of a synthesizer IC. For the register to recognize and prepare for the receipt of data, an ENABLE line must be raised before and during and then dropped at the completion of the data transfer as shown in Fig. 2. Following the first falling edge of the clock pulse after the drop of the ENABLE line, data will be transferred from the receiving register to the appropriate functions within the tuner.

The first four bits clocked into the register are band bits, which are essentially pass-through data that is used to enable the internal tuner band selectors (diode switches). Following the band bits in sequence are 14 or 15 frequency-select bits that are used to set up the synthesizer divider chain. These bits are in binary order with the MSD (most significant digit) bit being the first clocked in and the LSD (least significant digit) bit being the last. Some synthesizer registers will transfer the frequency data following the 19th clock bit should there be a data overrun. However, other tuners will fail to realize that the data has been overrun, and will ignore the entire entry. When an 18-bit count is used, many synthesizers will reset the first MSD (bit 15) frequency bit to zero while a 19-bit clock keeps bit 15 active. Any clock speed below 100 kHz is accepted by the synthesizer register, making it easy for manual data entry.

Data bits in the serial bit stream are given a binary weight, with bit 15 being equal to 16,384, bit 14 equal to 8,192, etc., down to bit 1, which is equal to one.

Voltage requirements for digital tuners appear to be quite uniform for all of those that I've encountered — including the I2C variety. The voltages are as follows: +12 V, +5 V, and +30 V. Maximum tuner sensitivity occurs with an AGC voltage between (+)6 V and 9 V.

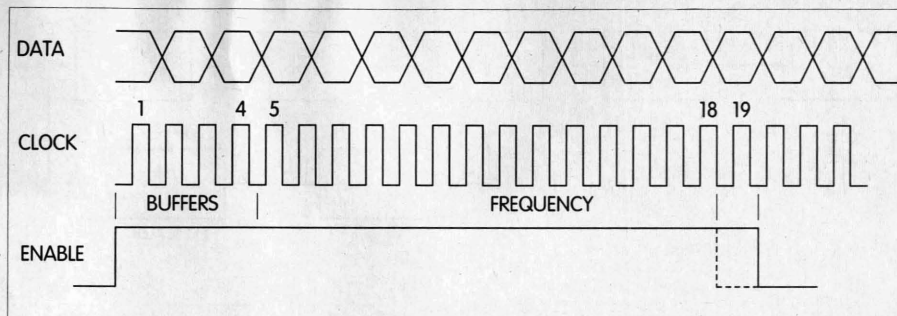


Fig. 2. 18/19-bit bus timing diagram for a 3-wire digital tuner.

Controlling the tuner is a matter of generating a serial bit stream that will select the desired frequency band within the tuner and then set up the synthesizer divider to the proper count. Determining the divider count is fairly straightforward and can be done with a pencil and paper, a calculator, or a computer. I've worked up a BASIC program to do the counting following the entry of a desired frequency. The BASIC program will be discussed later on, for those interested in using such a tool. I'll also show you how to easily work out the binary coding using the pencil and paper method.

After going through all of the above information, you may wonder why anyone would bother with a digital tuner. That's a good question that can be answered only by each individual. For me, the challenge of making the digital tuner operate was my reason for proceeding. Now that the digital tuner is operating, it has many of the same applications of the earlier varactor tuner.

Perhaps the best feature of the digital tuner is the frequency stability of the local oscillator (VCO). Being phase-locked, the local oscillator is as stable as the 4 MHz crystal that's used within the synthesizer. Defining all of the

possible uses for the tuner is a little difficult at this point, because only time and experimentation will provide the answer. However, the positive features involve frequency stability while monitoring a received signal, the input frequency is a "known" because of synthesizer control, and the tuner should make a useful and very stable synthesized signal source.

The front end design of the digital tuner incorporates a dual-gate FET, which allows it to have perhaps the best possible input sensitivity. During my tests, the input sensitivity has been as good as the receiver that followed the tuner. In some cases, the tuner provided an increase in input sensitivity. As an example, if the basic receiver exhibited an input sensitivity of 1 μ V, the use of the tuner allowed the sensitivity to remain at 1 μ V, and in some cases it improved to almost 0.5 μ V.

Perhaps, on the negative side, the digital tuner might not be a good candidate as the front end of a spectrum analyzer project because of the synthesizer step frequency. Unlike an analog varactor tuned tuner, which can be tuned smoothly from one frequency to another, a synthesized tuner will step

Continued on page 56

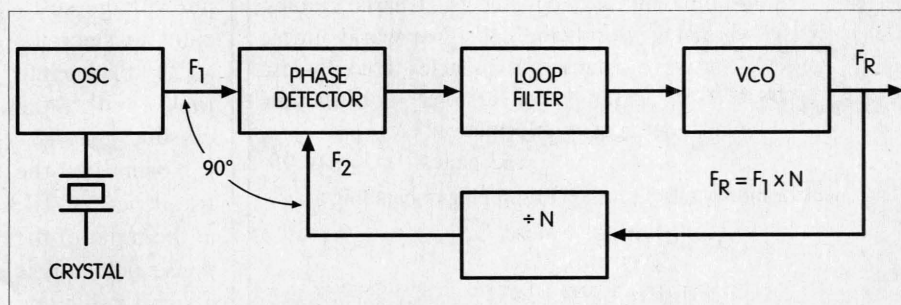


Fig. 3. Block diagram of a basic phase-locked loop synthesizer.

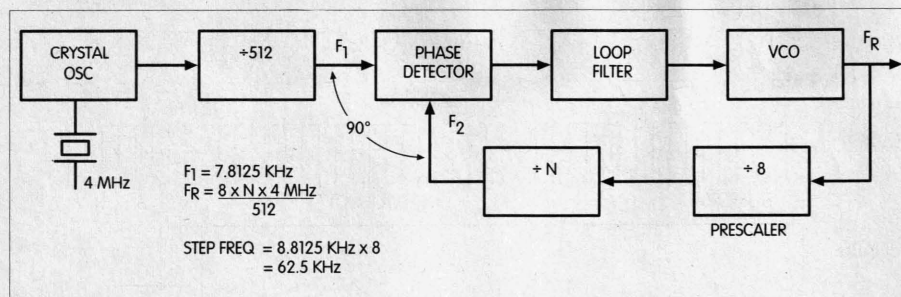


Fig. 4. Typical TV/VCR digital tuner synthesizer. Tuner step frequency is 62.5 kHz.

Inside Digital TV/VCR Tuners

continued from page 55

from one frequency to the next as the tuner moves across a frequency band.

Synthesizer theory

Before getting into the details of how the digital tuner is controlled, it is desirable to review how synthesizers operate, because that's the basis upon which a digital tuner operates. The objective is to understand how a synthesizer is used to control the local

oscillator (VCO) within the tuner through the use of a numerical divider.

Frequency synthesizers get their name from being able to generate a great number of discrete frequencies with the same stability as that of a quartz crystal, while utilizing only one crystal. Fig. 3 shows the basic elements of a phase-locked synthesizer where the VCO is locked to the stability of the crystal, but can operate at any frequency that can be locked by the loop. When the VCO is "locked," the two input signals to the phase detector are

equal in frequency, but operating in quadrature (90 degrees apart).

As long as the phase angle remains, the system operates OK, but if the phase shift changes between F1 and F2, then the VCO begins to move until either phase lock is again achieved or the VCO falls to a low preset out-of-lock frequency. In most cases, the phase detector will develop a DC output voltage sufficient to steer the VCO back into lock with the crystal.

Normally, the frequency of F1, at the input of the phase detector, is a small value and determines the

step frequency of the synthesizer. It is also used to calculate Fr (VCO output) when the loop divider is known. As an example, if F1 = 1 MHz and the divider is 10, then Fr = 10 MHz. A loop filter is used to reduce the VCO's rate of change so that "hunting" is eliminated.

A digital TV/VCR tuner synthesizer, as shown in Fig. 4, follows the theory described, but has two additional divider elements that modify the calculation. A prescaler (divide-by-8) is typically used between the VCO and the synthesizer divider. The purpose of the prescaler is to reduce the input frequency to the main divider because of the divider's low frequency response and the limited divide ratio that's available. The prescaler's fixed divide ratio has a direct effect on the final synthesizer step frequency.

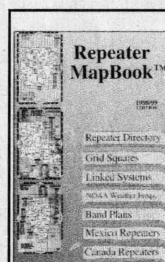
Between the crystal oscillator and the phase detector is an additional divider that reduces the crystal frequency to a low value. As shown, the divider has a divide ratio of 512, which reduces the 4 MHz crystal frequency to 7.8125 kHz (F1). With both fixed dividers in place, the equation for determining the step frequency is $7.8125 \text{ kHz} \times 8 = 62.5 \text{ kHz}$.

When stepping across the TV spectrum, a step frequency of 62.5 kHz is quite logical, but the step is pretty drastic for some ham applications. However, the wide step doesn't daunt a ham experimenter, because tuning across the tuner's IF output band with a receiver resolves the issue.

What's to follow

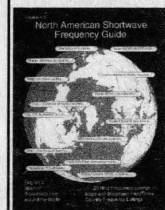
This is the first in a series of seven discussions on digital TV/VCR tuners. In this first section, we've discussed the different types of digital tuners and the frequency synthesizers used in the tuners, and have provided some insight into the calculations needed for digital tuner control.

The second part in the series will discuss the data transmitter that may be used for tuner control. Additional sections will include a data receiver, testing of the digital tuner, a BASIC program for generating the binary code, and a discussion on making printed circuit boards for a digital tuner project.



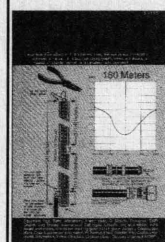
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Inside Digital TV/VCR Tuners

Part 4: Testing and binary data format.

The first three parts of this series on digital TV/VCR tuners discussed the two types of tuners, along with control and the test system that I used during my study of the digital tuner. Specific subjects covered were the synthesizer used to control the local oscillator (VCO) within the tuner, a data transmitter used for sending control data to the tuner, and a data receiver used for observing the data that has been sent to the tuner. In this part of the series we will discuss "how to test" the digital tuner and will describe the methods for setting up the binary data format used for controlling a digital tuner.

The testing of a digital tuner is reasonably simple once a method has been developed for clocking serial data into the tuner. With a data transmitter, microcontroller, or computer operating as a controller, it's necessary only to enter band select information and frequency, and select data in a binary format for the tuner to operate. Upon receiving data in the correct format, the tuner functions in a manner like that of the more familiar analog varactor TV/VCR tuner. Binary formatted data is used to control the tuner's frequency synthesizer. **Fig. 1** shows the data sequence format from MSD on the left to LSD on the right.

The frequency synthesizer within the digital tuner has a main divider chain that sets up the frequency ratio between the tuner's local oscillator (VCO) frequency and the reference frequency applied to the reference port of the phase detector. When the divide ratio number is known, it is then converted from a decimal number to a binary number. In a binary format, the ratio number is clocked serially into the tuner's register.

For the TV/VCR digital tuners that I have encountered, the main divide ratio

is determined by dividing the VCO frequency by 62.5 kHz. Some digital tuner synthesizers have optional step frequencies other than 62.5 kHz, but when used, the local oscillator (VCO) frequency band is altered to some degree. Gaining control of the optional step feature usually requires setting up the control data format to clock 34 bits of data instead of the normal 18/19 bits.

As an example of the process involved in finding the data format, choose a VCO frequency of 101 MHz, as is the case when the tuner is set for TV channel 2. Dividing 101 MHz by 62.5 kHz yields a divide ratio of 1616. Converting 1616 into a binary number provides the data format that can be clocked into the tuner's synthesizer. The binary conversion may be done using a great many different techniques, but the easiest is shown in **Fig. 2**. Although the use of a calculator will

speed the conversion process, it isn't required when the indicated steps are followed. To help keep track of the final number and to verify the accuracy of the conversion, it is recommended that a worksheet as shown in **Fig. 2(b)** be used as a guide.

Method for calculation

Starting with the known information such as the desired oscillator frequency and the synthesizer step frequency, determine the decimal divide ratio that is needed to generate the desired oscillator frequency.

When the divider ratio is known, subtract the largest binary number from it. In the example, the divide ratio is 1616 and the largest binary number that can be subtracted from it is 1024. Subtracting 1024 from 1616 leaves a remainder of 592. A binary "1" is assigned to the location in the

MSD																			LSD
4	3	2	1	15	14	13	12	11	10	9	8	7	6	5	4	3	2	1	
Band Selectors				Frequency Select Data in Binary Format															

Fig. 1. Data format required for controlling a digital TV/VCR tuner.

Known	Desired oscillator (VCO) frequency = 101 MHz															
	Synthesizer step frequency = 62.5 kHz															
Find	Synthesizer divide ratio															
	Binary number to be clocked into the tuner's register															
Steps	N = divider ratio = OSC freq/step freq = 101 MHz/0.0625 = 1616															
	(a)	(b)	(c)	(d)	(e)	(f)	(g)									
	1616 – 1024 = 592	592 – 512 = 80	80 < 256	80 < 128	80 – 64 = 16	16 < 32	16 – 16 = 0									
Assign	1	1	0	0	1	0	1									
(a)																
Osc. Freq.	N	16384	8192	4096	2048	1024	512	256	128	64	32	16	8	4	2	1
101 MHz	1616	0	0	0	0	1	1	0	0	1	0	1	0	0	0	0
448 MHz	7168	0	0	1	1	1	0	0	0	0	0	0	0	0	0	0
(b)																

Fig. 2. Shows the required input information and the steps involved in converting a decimal number to a binary number. (a) Shows the steps used to find and convert from a decimal number to a binary number. (b) Shows how the binary number is charted in the desired format for clocking data into the digital tuner's register.

chart under the number 1024. A binary "1" is assigned each time a number can be subtracted from the remainder. A binary "0" is assigned in the location when a subtraction cannot

be performed. In the case of the remainder "80" being smaller than "256," there is no subtraction so a binary "0" is assigned under the 256. Also in the example, 16 is the last

number where a subtraction takes place which allows the assignment of "0" to all of the other binary locations.

So that the process of converting from a decimal number to a binary number is clear, a second example is provided in the chart for a VCO frequency of 448 MHz. In this case, a synthesizer divide ratio of 7168 will be required. As the subtraction process occurs, a binary "1" is placed in the columns headed up by 4096, 2048, and 1024. All remaining columns contain a binary "0".

In part five of this series I'll provide a BASIC program that will calculate the binary number for any tuner "receiver" frequency, along with the local oscillator frequency that is selected.

Test setup

Fig. 3 shows the connections to the digital tuner, data transmitter, data receiver, and all of the power supply voltages. A voltage table is provided that may be used as a guide as to the typical current that the user should supply for operating all of the pieces contained in the test system.

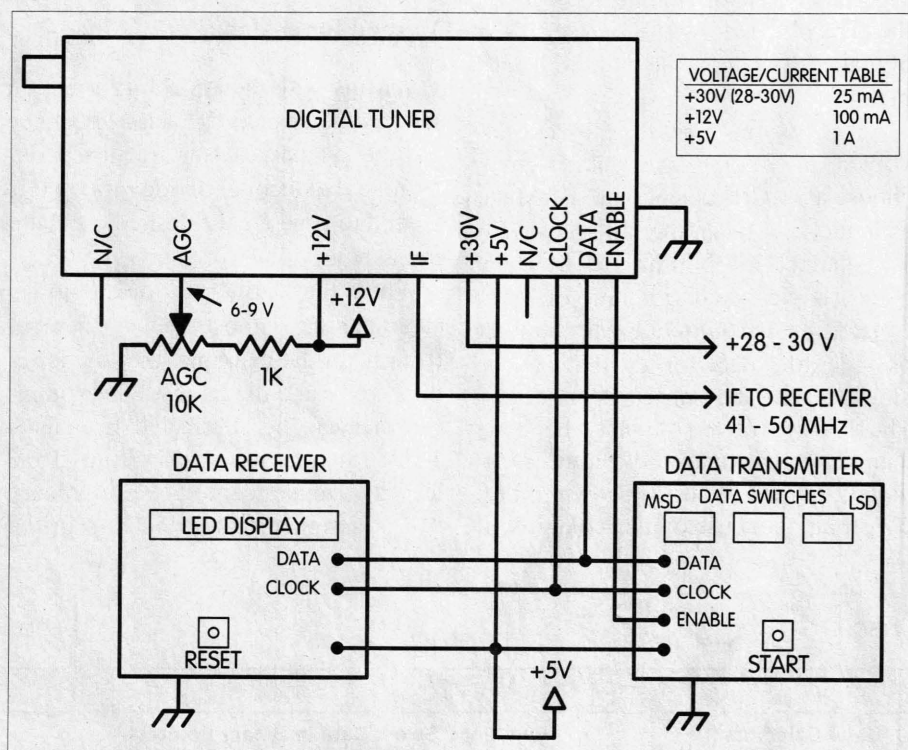


Fig. 3. Test setup and power for testing and/or using a 3-wire digital tuner.

Even though the figure is pretty clear as to how things are connected, a few words may help clear up any questions that might remain. All like terminal functions are tied together on all items. As an example, the data terminals on the tuner, data transmitter, and data receiver are connected together. The wires should be of short length and insulated, but shielding is not required. In other words, excess wire length should be avoided, but the units do not have to be crowded.

To gain a perspective on the physical layout of a digital tuner, refer to Fig 4. The two main areas of interest are the mixer/oscillator section, and the synthesizer section. The synthesizer section is readily identified because there will be a crystal mounted close by the synthesizer IC. In most tuners, the mixer/oscillator IC will be a surface mounted device soldered onto the circuit side of the circuit board (bottom side). Knowing the physical layout of the tuner becomes important during the test and checkout of the tuner.

Testing the tuner

All of the required voltages must be applied to the tuner so that it will be active and ready to operate. The input data is entered into the tuner by setting the switches on the transmitter and pressing the "start" switch. The tuner resets for the next data entry whenever the ENABLE line goes HIGH. Data entered previously into the tuner is retained by the tuner as long as power is applied and the ENABLE line remains LOW.

Band and frequency select data can be sent to the tuner at any time after power is applied. But during the initial stages of testing, the data set position relative to how the synthesizer data register "sees" the data can be an unknown and requires some initial experimentation. Shifting the data bank back and forth a bit or two will usually suffice, but finding the MSD and or LSD bit location within the tuner's register may be a little elusive.

One technique that I've used that appears to work with most tuners, particularly those having a synthesizer chip with known band control pinouts

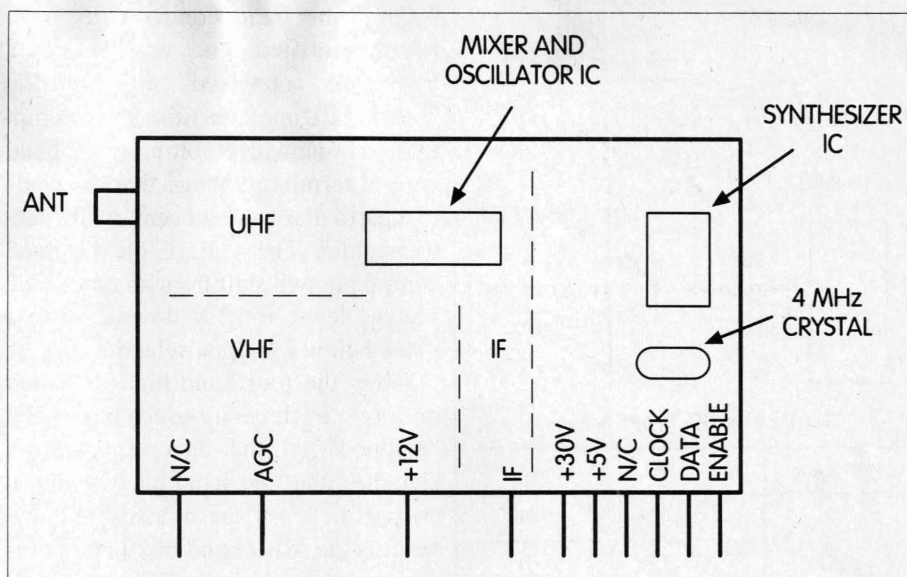


Fig. 4. Typical component placement and RF sections within a digital tuner.

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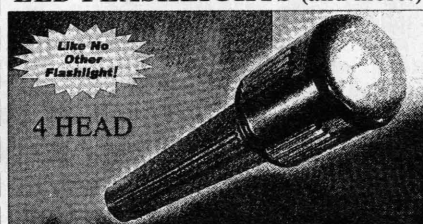
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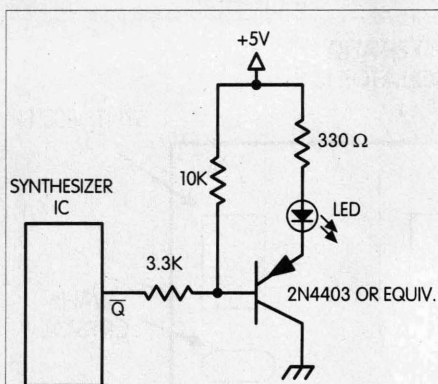
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Synthesizer IC Types & Pins

Motorola	MC44817D	Pin 15
	TD6359 P/N	Pin 20
Toshiba	TD 6380 P/N	Pin 20
	TD 6380 Z	Pin 16
	TD 6381 P/N	Pin 20
	TD 6381 Z	Pin 16
	TD 6382 P/N	Pin 20
	TD 6382 Z	Pin 16

Fig. 5. Synthesizer phase lock indicator. LED transistor driver is connected to the lock detector's output pin.

(one of four), is to measure the voltage on the selected pin. In the absence of known pinouts, tracing the tuner's band control transistor base circuits back to the synthesizer IC provides a level of confidence. A strong magnifying glass and an ohmmeter are usually required during the tracing process.

Once the band control pins have been identified, and with the data transmitter connected, enter only the "band" bits one at a time and attempt to determine with a voltmeter the band control terminal voltage that responds to a particular band bit sent by the data transmitter. The voltage on the band control pin will shift from an unselected voltage level of 12 V downward to a value below 7 V upon selection.

When the four band bits (of which only two or three are used) have been "mapped" on the data transmitter's switches, the first band bit (whether a "0" or a "1") to be clocked into the tuner is the MSD band bit, with the remaining three bits to follow. Because the band select bits are "pass-through," only one of four bits is selected for each of the tuner's bands.

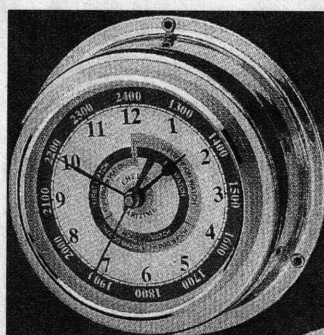
The first frequency data bit (MSD) will be the first data bit that follows the fourth band bit. The LSD data bit will be the last bit to be clocked into the synthesizer. An illustration of the band and data bit format that is expected by the digital tuner was shown in Fig. 1.

A more random method for determining the data set position for the data transmitter's switches is to find the lowest LSD switch setting that affects the synthesizer's divide ratio. Counting the switches upward

to 19 will identify the MSD position. With a step frequency of 62.5 kHz, the lowest LSD switch will shift the oscillator frequency by 62.5 kHz. The next lower switch will have no effect on the divide ratio.

One of the most helpful hints that I can provide is to suggest monitoring the synthesizer's "lock" feature. Entering data into the synthesizer and not knowing whether or not it's responding is quite unnerving at times. Most digital tuner synthesizer IC's have a dedicated pin that goes to a logic LOW when the system locks. Building up an LED driver circuit as shown in Fig. 5 will provide visualization of what the synthesizer is doing. Connecting the LED driver, as shown, requires that a wire be soldered to the appropriate IC pin (IC pins are indicated for specific chips). This step should be avoided if you lack skill in soldering in cramped spaces. Excessive heat must be avoided to prevent damage to the synthesizer IC.

Of course, monitoring the local oscillator frequency with a frequency counter, when one is available, will provide direct feedback as to what the synthesizer is doing as well as indicate



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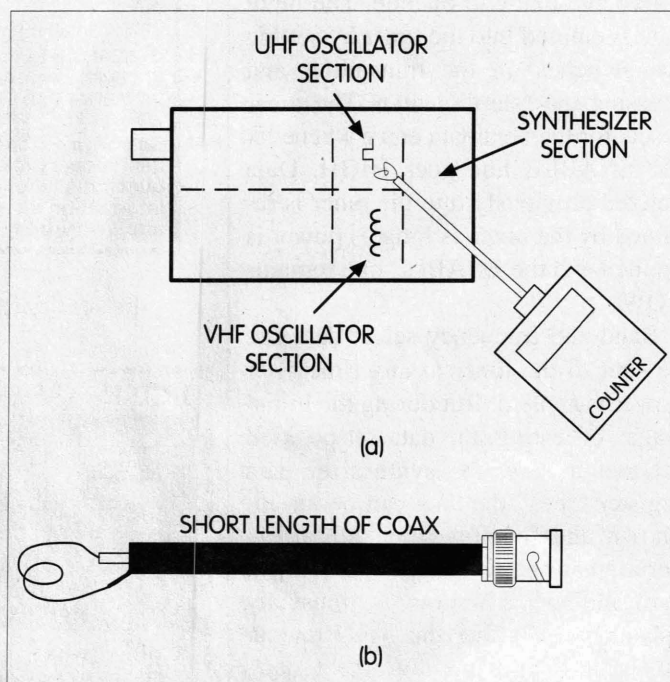


Fig. 6. Coupling signal energy from the tuner's VCO into a frequency counter. (a) Shows the approximate placement of the pickup coil. (b) Shows details of a suitable counter sampling probe. 2T insulated wire; coil diam. about 1/4"; coax type optional, RG-174, RG-58 work well.

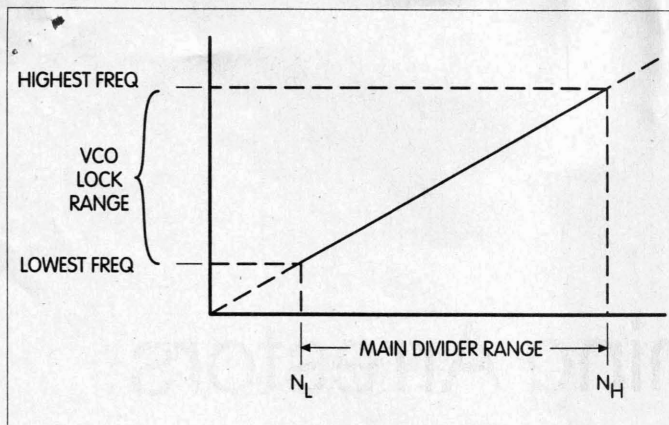


Fig. 7. VCO lock range as a function of the divide ratio of the main synthesizer divider. Dotted ends indicate ambiguity of the lock range per individual tuner.

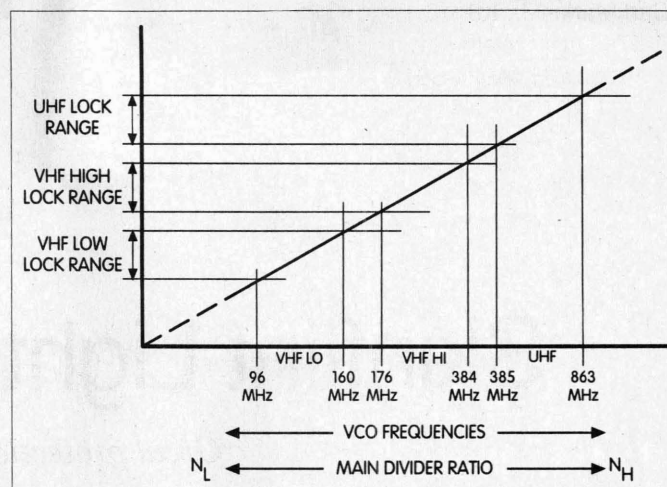


Fig. 8. Typical VCO tuning curve showing frequency vs. lock range.

the frequency of VCO operation. Coupling is provided to the counter through a small pickup loop that is placed adjacent to the oscillator coils within the tuner. Fig. 6 shows the details of the pickup loop and typical placement within the tuner. A tight coupling between the coupling loop and the oscillator coil is usually needed to achieve a "good" count.

Tuner response

One of the problems that I encountered during the initial test period was determining if and how the tuner might be responding. The use of the counter and "lock" indicator were of great assistance to me. Perhaps not knowing the band edges was the biggest deterrent.

To share my findings regarding the band limits and VCO lock capability, I've developed two charts shown in Figs. 7 and 8. The first chart is an expanded section showing the VCO tuning ramp from the lowest to the highest divide ratio for a given band. The VCO in various tuners has been set up to "lock" within the frequency requirements of the TV channels. But in some cases, the VCO will lock at a band of frequencies wider than the TV requirements, as indicated by the dotted lines representing the lock ambiguity. During initial testing, finding the near center frequency in each band provides the best opportunity of getting a "lock." Locating the lowest and highest frequency for each band is done by changing the divide ratio incrementally until the synthesizer drops out of lock.

Fig. 8 shows the typical VCO tuning curve and band by frequency. Some tuners exhibit a band gap between segments and others do not, which is a function of the lock ambiguity. The frequencies shown in the chart are the typical band limits that may be used for finding the near band center "lock" frequency during initial testing.

What's next

Parts five, six, and seven of this series on TV/VCR digital tuners will follow. Part five will provide a BASIC program that will allow the conversion of decimal frequency numbers to binary control numbers as required for tuner synthesizer control. Parts six and seven will wrap up the digital tuner discussion with a procedure for making printed circuit boards.

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It is good common practice to provide outer shield coax grounding at the base of a tower so that if a near miss lightning strike occurs, proper safety precautions are observed. However, there is the problem of the 2,000 volt breakdown between the inner conductor and shield of the coax cable. By the time the breakdown occurs, any

equipment left connected to it in the ham shack would have suffered substantial degrading, if not outright destruction. It is a costly problem for everyone concerned.

The first thing to be considered is the basic three classes of coax cable. The large 4,000 volt breakdown types, such as RG-213, RG-8, RG-17, and so

on, are relatively expensive in today's marketplace. Then there are the smaller cables used by hams who anticipate running output powers of 600 watts (continuous) or less. Quite a cost savings can be realized with such cable, along with the very flexible nature of the product. These cables are the 1000 volt breakdown types such as RG-58, RG-59, RG-8X, and so forth. Then there is the third type, such as RG-6, RG-174, special Teflon low loss, and so on. These types have breakdown voltage specifications of, typically, 700 volts and in some cases 200 volts. Obviously, these are to be used in the UHF/VHF ranges such as 145 MHz and 432 MHz; however, high SWR in the range of 3:1 can produce breakdown very easily in these types.

As can be seen, the buildup of EMP (electromagnetic pulse) between the inter and outer shield of coax can happen in just microseconds even with a near miss of lightning. If a strike *does* occur, it is probable that the tower will be the arrestor, and maybe you will not lose the rotor and rotor cable. Arresting of the rotor cable will occur at earth ground. Take the rotor cable and ensure that it is taped securely to the tower leg before descending down to

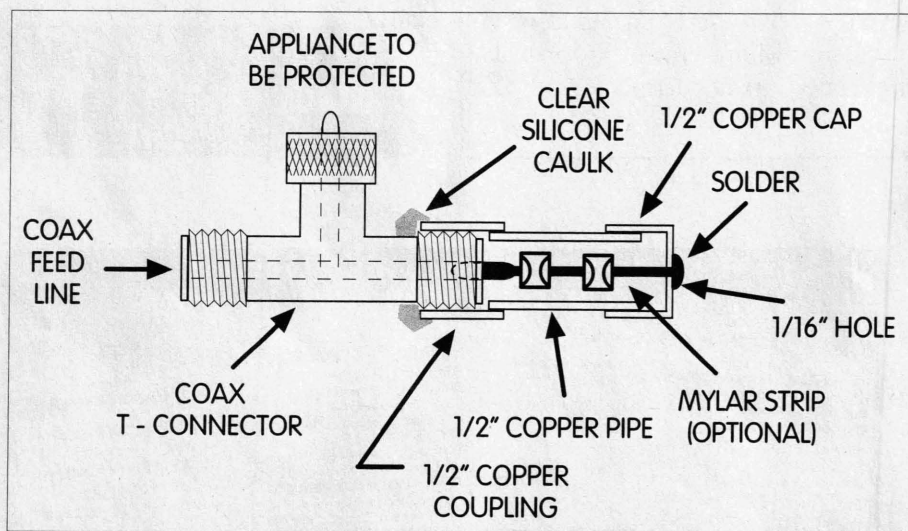


Fig. 1. Gas discharge coax lightning arrestor. Notes: (1) Ensure that gas devices do not short center conductor to shell. A piece of Mylar can be inserted. (2) Use clear silicone caulk around housing to make the moistureproof connection. (3) Two 350 V gas discharge spark gaps (700 V breakdown). Ensure centered, or use a piece of Mylar drafting material as an insulator (1 μ S 700 V response). (4) Specifications — 20,000 A surge current, 10^{10} ohm insulation resistance, 1 pF capacitance, 1 mS response (100 V/ μ S).